Frequency Shifter For "Howl" Suppression

*M. Hartley Jones**, *M.Sc., Ph.D.*

The circuit described increases the stability margin of sound reinforcement systems by up to 8dB. This is achieved by shifting the frequency of signals by 5Hz. The inexpensive and reliable nature of the circuit is made possible by the use of integrated circuits. The technique may be adapted to produce larger frequency shifts for use in electronic music synthesis.

Anyone who is involved with sound reinforcement systems is familiar with acoustic feedback. A sound amplification system where the microphone and speaker are in the same room, is limited in gain by the positive feedback loop formed by the acoustic coupling between loudspeaker and microphone. As the gain is increased, feedback initially causes colorations in the sound, then develops into an audible ringing on transients, finally becoming a continuous "howl-round" when the loop gain reaches unity.

The use of directional microphones and loudspeakers can help in reducing the proportion of sound energy which is fed back to the microphone but, owing to the diffuse nature of the reverberant sound field of a room, acoustic feedback still limits the usable gain of the system.

The reason why feedback usually occurs at one particular frequency is an interesting one. In any sound reinforcement system there will be one frequency at which loop gain is a maximum and this will depend upon the frequency response characteristics of the complete feedback chain, comprising microphone, amplifier, loudspeaker and room, as shown in Fig. 1. If frequency response peaks can be ironed out of every element in the chain, then the mean loop gain can be increased without exciting howl-round at a particular frequency.

Compensation for frequency response irregularities in microphones and loudspeakers by means of tone controls and equalizers in the amplifier is a well-established technique and attention to this point can often improve the stability margin of a system by several decibels. The frequency response of the room is, however, a much more complicated problem owing to the multiplicity of resonant modes. Fig. 2 shows the typically irregular frequency response of a room; only the region just above 500Hz is shown, but the characteristic over the whole audio band is of similar form. The dotted line in Fig. 2 shows the average or mean level of response and it is clear that the peak levels can be more than 10dB above the average. Although it is the average response which determines the subjective sound level in a room, as judged by a listener, it is those 10dB peaks that give rise to howl-round in a sound reinforcement system. Thus, if the peaks could be flattened out, up to 10dB increase in usable gain could be expected.

There have been attempts at electrical equalization of the major room response peaks[1] but, even if the inverse response of Fig. 2 could be exactly synthesized, the peaks and dips shift with variations in microphone position, with temperature and with size of audience, making exact correction impossible.

Schroeder[2,3] has shown that the room frequency response can be effectively smoothed out as far as acoustic feedback is concerned by slightly shifting the frequency

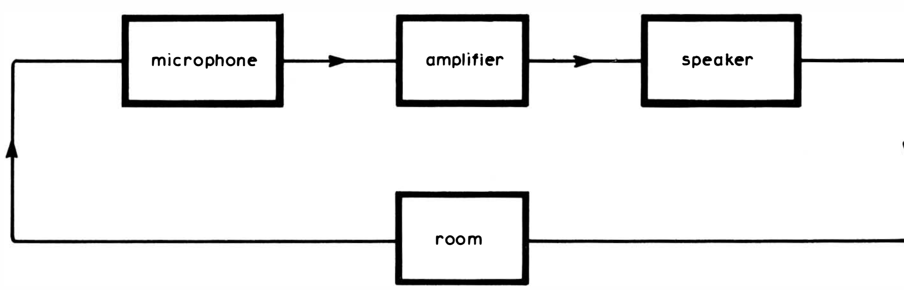

Fig. 1. Acoustic feedback loop in a sound system.

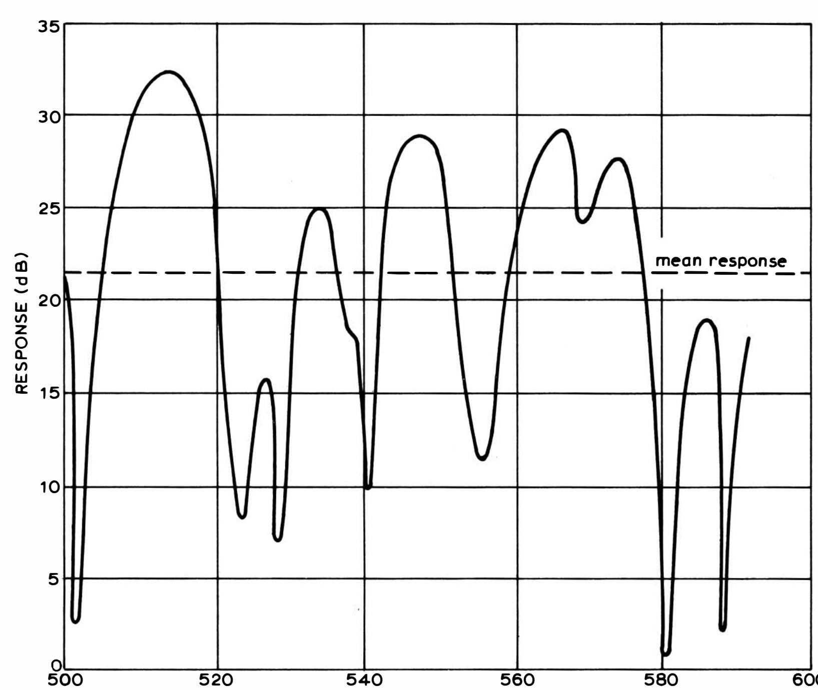

Fig. 2. Small section of a typical frequency response curve of a room.

*University of Manchester Institute of Science and Technology.

of the signal as it goes through the amplifier. This technique ensures that any signal component subject to a very high gain at one of the response peaks, which would normally result in a feedback howl, will be changed in frequency on its next trip round the loop so that it is subject to a much lower gain. As the signal continues to go round the loop it will be changed in frequency each time and therefore be subject to a different loop gain on each trip. After several trips, the mean loop gain experienced by the signal will thus be equal to the average room response shown in Fig. 2. In the same way, if a signal component happens to experience a low gain on its first trip then it will be subject to a much higher gain on its second trip. However, after several trips, the mean loop gain will once again be equal to the average room response. Thus both peaks and dips in the room response are effectively levelled by the frequency shifter.

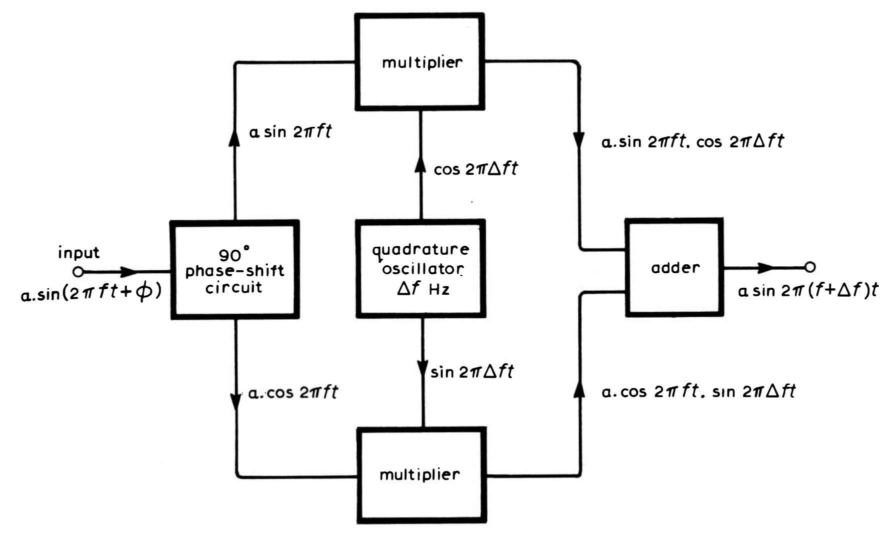

Fig. 3. Block circuit diagram of the frequency shifter.

The amount by which the frequency is shifted is not particularly critical. In order to minimize unnatural effects the shift should be as small as possible, whilst for optimum howl suppression the shift should be of similar order to the average spacing between peaks and their adjacent troughs. Schroeder and Kuttruff[4] have shown theoretically that the average spacing between adjacent peaks is given by:

$$\Delta f_{max} \approx \frac{4}{T}\,\text{Hz}$$

where T is the reverberation time of the room. Thus, on average, in a room with a reverberation time of one second, the spacing between adjacent peaks should be 4Hz and therefore that between a peak and its adjacent trough 2Hz. This average spacing, however, includes tiny irregularities which may be barely measurable; in practice the spacing of major peaks and troughs may be double this figure; in addition, as may be seen in Fig. 2, the peaks are not regularly spaced. The optimum frequency shift for

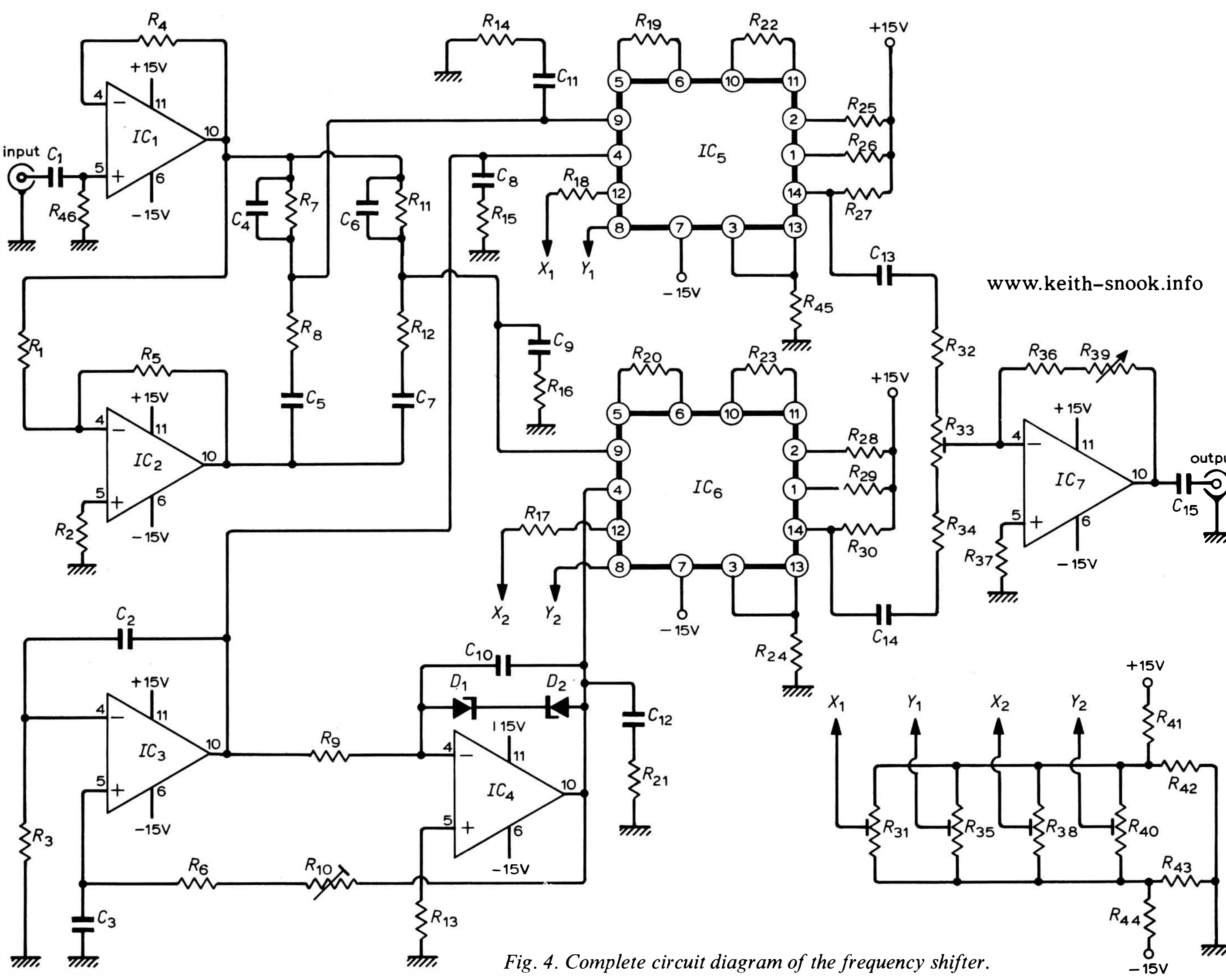

Fig. 4. Complete circuit diagram of the frequency shifter.

howl suppression is therefore best found experimentally; in most applications it turns out that a shift of 5Hz is sufficient, no further advantage being gained by greater shifts. It is conceivable that, in the unlikely event of the reverberation time being very much less than one second, a greater shift might be required.

It is immaterial whether the frequency shift is in the upward or downward direction. The shifter to be described produces a 5Hz increase in frequency.

Frequency shift circuit

Schroeder used in his experiments a frequency shift circuit developed by Prestigiacomo and MacLean[5]. This circuit used suppressed carrier single-sideband modulation of a high frequency carrier by the signal followed by demodulation with a new carrier frequency which differed from the original carrier by 5Hz. The circuit to be described here, however, uses a different technique, exploiting the accurate integrated circuit multipliers now freely available; it requires no tuned circuits, crystals or filters and gives a good signal to noise ratio with a wide frequency response and low distortion.

Consider a sinusoidal input signal of frequency fHz; its instantaneous voltage, v_{in}, is represented by

$$v_{in} = a \sin 2\pi f t \qquad (1)$$

where a is peak voltage.

Now, if the output signal of our circuit has an increased frequency $(f+\Delta f)$Hz but the same peak voltage a, then its instantaneous amplitude v_{out} is

$$\begin{aligned} v_{out} &= a \sin 2\pi(f+\Delta f)t \\ &= a \sin(2\pi f t + 2\pi \Delta f t) \qquad (2) \end{aligned}$$

Expanding this sine function gives:

$$v_{out} = a(\sin 2\pi f t \cos 2\pi \Delta f t + \cos 2\pi f t \sin 2\pi \Delta f t) \qquad (3)$$

Equation 3 is synthesized by the circuit to be described. The cosine functions in the equation are simply sine functions with a $\pi/2$ (90°) phase advance imposed by a phase shift circuit, since $\cos x = \sin(x+(\pi/2))$. Although equations 1 to 3 refer specifically to sinusoidal signals, they may be applied to complex waveforms by Fourier analysis into sine and cosine components.

Fig. 3 shows the basic circuit elements of the frequency shifter. Sinusoidal and cosinusoidal components are derived from the original signal by a broad-band phase shift circuit. An oscillator with quadrature outputs generates sinusoidal and cosinusoidal components of the "shift frequency" Δf. Two analogue multipliers give the products $\sin 2\pi f t . \cos 2\pi \Delta f t$ and $\cos 2\pi f t . \sin 2\pi \Delta f t$ and an operational adder gives the final frequency shifted signal:

$$\sin 2\pi f t . \cos 2\pi \Delta f t + \cos 2\pi f t . \sin 2\pi \Delta f t = \sin(2\pi f t + 2\pi \Delta f t).$$

Circuit description

The complete circuit diagram is shown in Fig. 4. The frequency shifter is designed to give unity voltage gain, but an extra 4dB voltage gain is available if required. The

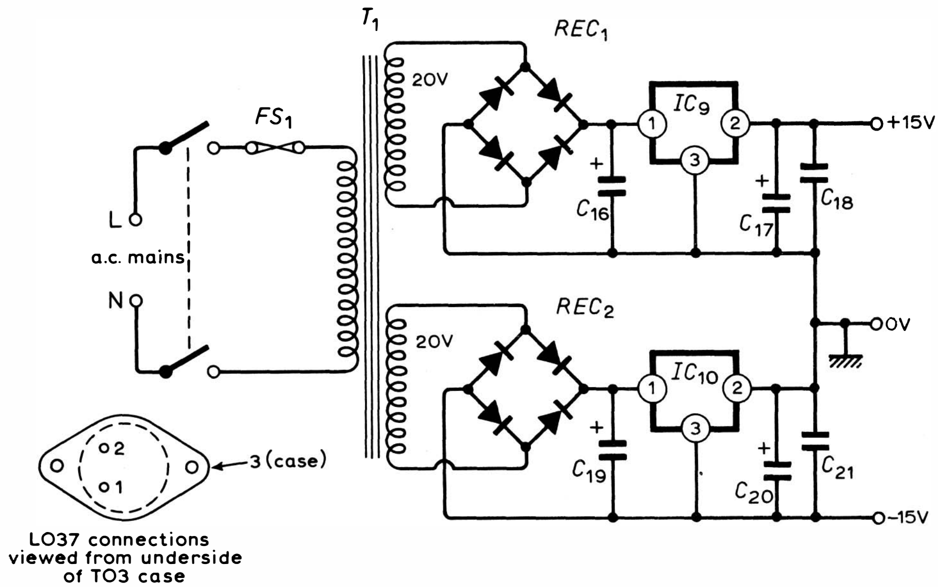

Fig. 5. Power supply circuit.

maximum input voltage before clipping occurs is 4V.r.m.s. (+14dBm, where 0dBm = 0.775V.r.m.s.). The unit is thus suitable for connection between the microphone pre-amplifier and main power amplifier of a sound system. A modification is described later which makes the circuit suitable for use directly in a microphone line.

The input signal feeds first into a voltage follower, IC_1, giving a high impedance input to the unit. The broad-band phase shifter uses IC_2 as an inverting amplifier and RC networks R_7, R_8, C_4, C_5 and R_{11}, R_{12}, C_6, C_7 to give two quadrature outputs of equal amplitude. The operation of this circuit is discussed in the appendix.

The sinusoidal and cosinusoidal components of the shift frequency Δf are generated by a quadrature oscillator incorporating amplifiers IC_3 and IC_4. This operates at 5Hz with the component values shown and gives sinusoidal and cosinusoidal outputs of equal amplitude and with less than 1% distortion. The 6.8V zener diodes across C_{10} act as amplitude limiters, the final adjustment of output amplitude being performed by R_{10}.

Analogue multipliers, IC_5 and IC_6, give the products of the input signal components with the shift frequency components. RC networks C_{11} and R_{14}, C_8 and R_{15}, C_9 and R_{16}, C_{12} and R_{21} are connected directly from the multiplier inputs to earth to suppress any tendency towards h.f. instability. Variable resistors R_{31}, R_{35}, R_{38} and R_{40} are the multiplier input offset controls which are adjusted during the setting up procedure to ensure that when one of the inputs of a multiplier is at zero volts then the output is also zero. Capacitors C_{13} and C_{14} on the multiplier outputs block the d.c. component inherent in the output of this type of multiplier chip and couple the a.c. products to the final stage.

IC_7 is used in the final stage as an operational adder with R_{39} in the feedback loop to adjust overall gain. Variable resistor R_{33} is a balance control to compensate for any amplitude difference in the products from the two multipliers.

IC_1, IC_2, IC_3, IC_4 and IC_7 are all type 741 internally compensated operational amplifiers. The Texas Instruments SN72741N 14-pin dual-in-line version was used in the prototype. IC_5 and IC_6, the analogue multipliers, are type MC1495L (Motorola) or SG1495D (Silicon General). All the pin connection numbers on the circuit diagram refer to 14-pin dual-in-line packages.

Resistors are all $\frac{1}{8}$W, ±5% tolerance. Capacitor tolerances are all ±20% with the exception of C_4, C_5, C_6 and C_7 which should be ±5% or better.

Power supplies

All the integrated circuits in Fig. 4 operate from supplies of +15V and −15V; the power supply circuit is shown in Fig. 5. Fixed voltage integrated circuit regulators (SGS LO37 or RS Components MVR 15V) are used to produce stable ±15V supplies from a conventional transformer and bridge rectifier arrangement. In the prototype, T_1 was an RS Components 20V miniature transformer and REC_1 and REC_2 were RS Components REC41A. C_{18} and C_{21} are 0.1μF polyester or ceramic capacitors to ensure a low supply output impedance at high frequencies.

Construction and adjustments

Circuit layout has not been found critical. The circuit was assembled on a piece of 0.1in matrix Veroboard of size 165 × 95mm and housed together with the power supply, which was on a separate matrix board, in a diecast aluminium box of external dimensions 190 × 120 × 50mm.

In the prototype unit, R_{39}, the gain control, was the only front panel potentiometer. Resistors R_{10}, R_{31}, R_{35}, R_{38}, R_{40}, and R_{33} were all skeleton presets mounted on the circuit board and, once these were correctly adjusted, the settings were found to be extremely stable and required no further adjustment. If no external gain control is required, R_{39} could also be pre-set.

When setting up the circuit, an oscilloscope and sine-wave audio generator are of great assistance but, if necessary, the adjustments can be made using an a.c. voltmeter (e.g. AVO model 8) and a sinusoidal signal from radio, TV or record combined with careful listening. Before switching on it is advisable to set all potentiometers to their mid-position; most of them will then be fairly near the correct setting. The best order of adjustment is as follows:

- The 5Hz sine wave amplitude should be adjusted with R_{10} to 9 volts peak to peak, measured on an oscilloscope at pin 10 of IC_4. Approximately the same amplitude, but 90° out of phase, should be present at pin 10 of IC_3. If an oscilloscope is not available for this adjustment, an a.c. voltmeter (1000Ω/V) may be used, the output being set to 3V.r.m.s. This latter method has proved sufficiently accurate despite some oscillation of the meter pointer due to the low frequency.
- With no signal on the input, R_{31} and R_{38}, the multiplier X-input offset controls, are adjusted for minimum 5Hz component in the output. An oscilloscope or electronic millivoltmeter is essential for precise adjustment; however, if the controls are simply left at their mid-position, the 5Hz rejection will be sufficient for most applications. The response of most power amplifiers is much reduced at such low frequencies.
- With a sinusoidal signal of approximately 1kHz frequency on the circuit input, R_{33}, the adder balance, and then R_{35} and R_{40}, the multiplier Y-input offsets, are adjusted for minimum amplitude modulation in the output, viewed on an oscilloscope. This adjustment may, if necessary, be performed fairly accurately on a listening test using a continuous input, such as a radio or TV tuning signal and adjusting for minimum fluctuation in sound output from a power amplifier and speaker connected to the frequency shifter. A random noise signal (e.g. an f.m. tuner between stations) can also be used for a listening test adjustment.

Although R_{33}, R_{35} and R_{40} do not interact, it may be necessary to go through the adjustment procedure two or three times to obtain optimum results. Once the controls are set, no further adjustment should be required.

Measured prototype performance

In the measurements given below, the overall voltage gain of the frequency shifter was set to unity.

Frequency increase
5.3Hz

Max. input signal prior to clipping
4.0V r.m.s.

Noise level ref. 4V r.m.s. output (measured over a band from 30Hz–10kHz)
−73dB

Residual 5Hz component in output ref. 4V r.m.s.
−66dB

Frequency response
1dB down at 85Hz and 19kHz
3dB down at 45Hz

Residual amplitude modulation (applicable at all output levels)
<0.5 dB peak to peak output signal fluctuation from 250Hz to 3kHz
<1dB fluctuation from 220Hz to 13kHz

Total harmonic distortion measured at 1kHz
<0.1% at 3.8V r.m.s. output
<0.01% at 500mV r.m.s. output

Modification for microphone line

The basic frequency shifter design can be used in any medium or high impedance circuit, but, on a low-level microphone line, noise can become obtrusive. Most of the noise originates in the multipliers; hence the signal to noise ratio can be improved at the expense of overload margin by increasing the gain on the input side and attenuating the output by the same factor.

The simple circuit modification shown in Fig. 6(a) increases the gain of the input amplifier from unity to 27dB by reducing the negative feedback. The voltage divider shown in Fig. 6(b) then provides 27dB attenuation on the output to maintain unity gain overall. After the modification the maximum signal input before clipping is reduced to 170mV, but the noise level is reduced by 27dB. This version of the circuit has given good results in a microphone line with negligible noise increase and adequate overload margin. The unit has also been successfully used in a 30Ω balanced line with the addition of 30Ω to high impedance microphone transformers giving a step up at the input and a step down back to balanced line at the output.

Practical applications

It has been found in practice that the use of the frequency shifter in an average sound system makes available an extra 6 to 8dB of usable gain. Although it was indicated previously that the theoretical gain increase may be over 10dB, this figure is not realized in practice because the feedback that does eventually develop in a frequency shifted system has an unnatural "warbling" character which must be avoided. However, this unusual feedback characteristic has a positive advantage in that a sound system becomes easier to set up and operate because the onset of feedback is so readily detected. When the warbling becomes audible, there is normally an extra 3 to 4dB gain still in hand before instability develops, unlike the unpleasant "ringing" colorations heard in a conventional system which often occur only 1 to 2dB below howl-round.

Even when a conventional system is operating well below howl-round, the colorations due to acoustic feedback are often sufficient to degrade the sound reproduction so that high quality equipment is unable to give of its best. In these circumstances, the use of the frequency shifter has been found to give a positive improvement in sound quality, even though extra gain may not be required.

It should be emphasized that the circuit obtains its increase in stability margin by effectively flattening the frequency response of the room. In other words, it provides the greatest improvement when the sound from the loudspeaker arriving back at the microphone is of a reverberant character; i.e. when it has been "processed" by multiple reflections within the room. In most systems, even in relatively dead rooms, this is automatically the case; however, if the microphone is only a short distance from the loudspeaker (2–3 metres), then the available gain increase may be reduced to 2 or 3dB.

The frequency shifter has been used successfully not only with speech but also with music amplification. It is perhaps surprising that the 5Hz shift is not obvious, but the only effect noticed has been a slight 5Hz beat between direct and amplified sound on an occasional long continuous note. The general reduction of coloration in the amplification system with the frequency shift in operation outweighed any possible degradation of quality due to beating effects.

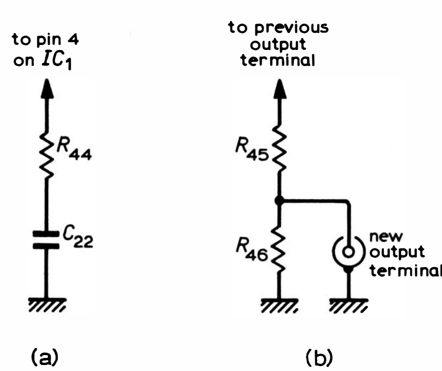

Fig. 6. Circuit modifications for low-level operation: (a) addition to feedback loop of IC_1; (b) output attenuator.

Electronic music

Although the circuit as described produces a fixed increase in frequency of 5Hz, the frequency of the modulating quadrature oscillator can be increased by reducing time constants R_3C_2, R_6C_3 and R_9C_{10} to produce shifts of several hundred hertz or more. As can be imagined, this shifting of a whole band of frequencies introduces weird effects which can be applied to electronic music; all normal harmonic relationships are destroyed because every frequency is shifted by a fixed number of hertz rather than being subject to a given fractional change.

Bode and Moog[6] have recently described a frequency shifter using the same principle as described here, but specifically designed for electronic music; the modulating frequency is generated by a voltage-controlled beat-frequency oscillator giving a continuous range of frequency shift from −5kHz to +5kHz. By using a subtraction circuit as well as the adder at the output, an upward shift and downward shift may be obtained simultaneously from separate outputs for unusual stereophonic effects. There is obvious scope for the modification of the present circuit along these lines for experiments in electronic music synthesis.

Conclusion

The use of a 5Hz frequency shift in a sound reinforcement system can dramatically reduce acoustic feedback, improving stability margin by 6 to 8dB. Although the technique has been known for over ten years, it has not yet achieved the widespread use it deserves. The circuit described gives high quality results and is inexpensive to build; a simple modification can provide large shifts of frequency applicable to special effects in electronic music synthesis. The total cost of the prototype was under £20.

Appendix

Design procedure of the broad-band phase shifter. The frequency shifter is a specialized example of the type of single-sideband modulator which depends for its operation on a broad-band phase shifter producing two components of the audio signal with a 90° phase difference. Such phase shifters are also appearing on the domestic scene in quadraphonic decoders; the design criteria of such circuits are therefore of some interest.

The principles of broad-band phase shifter design have been discussed by Dome[7] and Luck[8] and analyzed in detail by Orchard[9]. Stein[10] has published calculations on an *RC* lattice phase shifting network similar to the type used here.

To obtain two outputs with a constant phase difference over a wide frequency range, the principle used is to derive the outputs from two separate networks each of which produces a frequency-dependent phase shift but where the phase difference between the outputs is constant over the required range. Fig. 7 shows such a circuit in block form and Fig. 8 the type of phase/frequency relationship that will give the required result. It is desirable that the output amplitude from each network should be independent of frequency, i.e. that the network be of the all pass type.

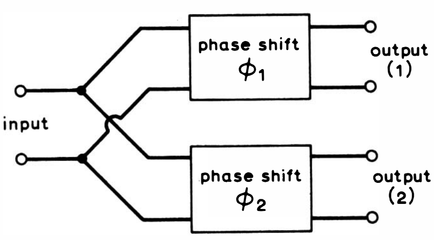

Fig. 7. Block diagram of a broad-band phase shifter.

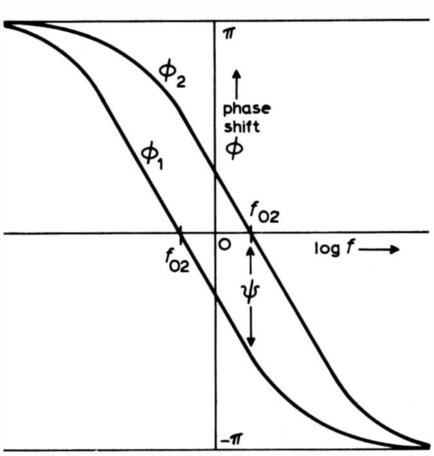

Fig. 8. Graph of phase shifts of individual branches of broad-band phase shifter plotted against frequency.

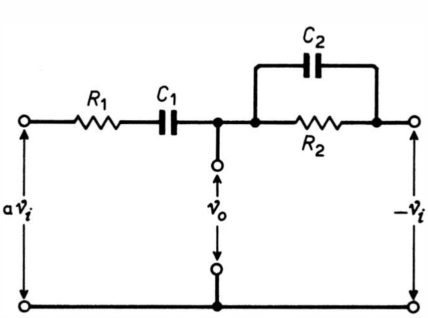

Fig. 9. One branch of the phase-shift network used in the frequency shifter.

Dome[7] has shown that a constant phase difference may be realized if the phase shift angle of each network varies according to the logarithm of frequency.

i.e. if $\phi_1 = C_1 + \log K_1 f$

and $\phi_2 = C_2 + \log K_2 f$

where C_1, C_2, K_1, K_2 are constants and ϕ_1 and ϕ_2 the phase shifts of the two networks, then phase difference $\psi = \phi_2 - \phi_1$

$$= C_2 - C_1 + \log\frac{K_2}{K_1}$$

$$= \text{constant.}$$

Luck[8] has indicated that the desired characteristic may be obtained with a pair of resonant circuits. In the phase angle graphs of Fig. 8, f_{o1} and f_{o2} are the resonant frequencies of the two networks. The required Q-value is very low (<0.5); *RC* networks can therefore be used. Each network used here is of the Wien type shown in Fig. 9. The following calculation shows that this type of network can fulfil the requirements for a broad-band phase shifter.

In Fig. 9, consideration of circuit impedances gives:

$$\frac{(a+1)v_i}{v_o+v_i} = \frac{C_2}{C_1}+\frac{R_1}{R_2}+1 + j\left(R_1\omega C_2 - \frac{1}{R_2\omega C_1}\right) \quad (1)$$

$$= \alpha + j\beta$$

giving $$\frac{v_o}{v_i} = \frac{a+1-(\alpha+j\beta)}{\alpha+j\beta}$$

Now if constant a is chosen so that:

$$a = 2\alpha - 1$$

then $$\frac{v_o}{v_i} = \frac{\alpha-j\beta}{\alpha+j\beta} \quad (2)$$

$$\therefore \sum\frac{v_0}{v_1} = \sqrt{\frac{\alpha^2+\beta^2}{\alpha^2+\beta^2}}$$

$$= 1$$

Thus the network is of the all-pass type. Now, when $\omega = \dfrac{1}{R_1C_1} = \dfrac{1}{R_2C_2}$,

$\beta = 0$ and the circuit exhibits resonance.

$$\text{Resonant frequency}, f_o = \frac{1}{2\pi R_1 C_1}$$

$$= \frac{1}{2\pi R_2 C_2} \quad (3)$$

It can now be shown that the relationship between phase angle and frequency is of the type required for a broad-band phase shifter.

Substitution for α and β in (2) together with f_o from (3) gives:

$$\frac{v_o}{v_i} = \frac{1+jQ\left(\frac{f_o}{f}-\frac{f}{f_o}\right)}{1-jQ\left(\frac{f_o}{f}-\frac{f}{f_o}\right)} \quad (4)$$

where $$Q = \frac{1}{\frac{R_2}{R_1}+1}$$

and has its usual significance as the quality factor of the resonant circuit. ($Q = f_o/\delta f$, where δf is the half-power bandwidth.)

From (4) the phase angle, ϕ, between v_o and v_i is obtained:

$$\tan\frac{\phi}{2} = Q\left(\frac{f_o}{f}-\frac{f}{f_o}\right) \quad (5)$$

For small values of ϕ, it may be shown[10] that:

$$\phi \approx -4Q\ln\frac{f}{f_o}$$

which is the type of logarithmic relationship required for a broad-band phase shifter. Two of these networks, of resonant frequencies f_{o1} and f_{o2} may therefore be used to produce a constant phase difference,

$$\psi = \phi_2 - \phi_1.$$

Luck[8] has calculated ψ by applying equation (5) to both networks and has deduced a set of design parameters relating Q, and f_{o1} and f_{o2} with the frequency range over which ψ is to be close to 90°. The greater the permissible deviation from 90°, the wider the frequency range which can be handled.

In the frequency shifter a range from 250Hz to 12kHz is satisfactory and Luck's results indicate that a 90° shift over this range should be possible with maximum deviations of $\pm 5°$. The necessary circuit parameters are:

$$f_{o1} = 720\text{Hz}$$

$$f_{o2} = 4{,}000\text{Hz}$$

$$Q = 0.21$$

thus,

$$\frac{R_2}{R_1} = \frac{C_1}{C_2} = 2.7$$

and $a = 2.5$

The circuit impedances chosen are sufficiently high to avoid excessive loading on the outputs of the drive amplifiers and yet low enough for the 20MΩ input impedance of the multipliers to have a negligible effect on performance. Theoretically the phase shifters should see an infinite load impedance.

Fig. 10 shows the measured phase shift performance of the practical circuit. A phase difference of $90° \pm 7°$ is obtained from 200Hz to 15kHz.

Deviations from 90° in the phase-shift circuit produce amplitude modulation in the output of the frequency shifter; this occurs at twice the shift frequency owing to the generation of an unwanted sideband. The effect only becomes significant below 200Hz and above 15kHz and is not sub-

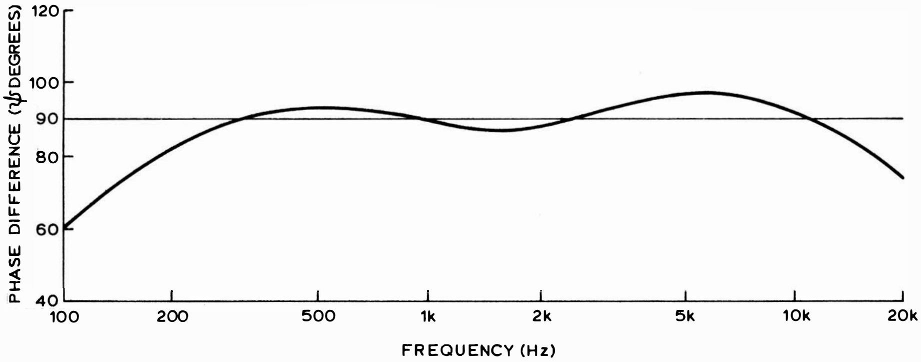

Fig. 10. Graph of measured phase difference plotted against frequency for broad-band phase shift network used in frequency shifter.

jectively troublesome in practice. A closer approximation to a 90° phase difference over a wider frequency range may be obtained if necessary by replacing each Wien network by several such networks in cascade with staggered resonant frequencies.

Components list

Resistors

R_1	33kΩ
R_2	22kΩ
R_3, R_4, R_6, R_{46}	270kΩ
R_5, R_{12}	82kΩ
R_7	43kΩ
R_8	15kΩ
R_9, R_{13}	330kΩ
R_{10}, R_{33}	100kΩ preset pot.
R_{11}, R_{17}	220kΩ
R_{14}, R_{15}, R_{16}, R_{21}	470Ω
R_{18}	39kΩ
R_{19}, R_{20}, R_{22}, R_{23}	8.2kΩ
R_{24}	6.8kΩ
R_{25-30}	3.3kΩ
R_{31}, R_{35}, R_{38}, R_{40}	10kΩ preset pot.
R_{32}, R_{34}	100kΩ
R_{36}	560kΩ
R_{37}	1MΩ
R_{39}	500k preset or panel mounting pot.
R_{41}, R_{44}	10kΩ
R_{42}, R_{43}	3.9k

All resistors $\frac{1}{8}$W $\pm 5\%$ tolerance

Capacitors

C_1, C_2, C_3, C_{10}, C_{13}, C_{14}	100nF
C_4, C_6	1nF
C_5, C_7	2.7nF
C_8, C_9, C_{11}, C_{12}	10pF
C_{15}	470nF
C_{16}, C_{19}	1000μF 40V electrolytic
C_{17}, C_{20}	10μF 16V electrolytic
C_{18}, C_{21}	100nF

Tolerance $\pm 5\%$ or better

Integrated circuits, Semiconductors

IC_{1-4}	Texas SN72741N
IC_5, IC_6	Motorola MC149SL or Silicon General SG1495D
IC_9, IC_{10}	SGS LO37
D_1, D_2	6.8V zener 500mW
REC1, *REC2*	REC41A RS Components
Transformer	TI 20V Min. transformer RS Components

References

1. Boner, C. P. and Boner, C. R. "Minimizing feedback in sound systems and room-ring modes with passive networks", *J. Acoust. Soc. America*, Vol. 37, p. 131 (1965).
2. Schroeder, M. R. "Improvement of acoustic feedback stability in public address systems", *Proc. Third Internat. Congr. Acoustics* (1959), Vol. II, p. 771, Elsevier Publishing Co., Amsterdam (1961).
3. Schroeder, M. R. "Improvement of acoustic feedback stability by frequency shifting", *J. Acoust. Soc. America*, Vol. 36, p. 1718 (1964).
4. Schroeder, M. R. and Kuttruff, K. H. "On frequency response curves in rooms. Comparison of experimental, theoretical and Monte Carlo results for the average frequency spacing between maxima", *J. Acoust. Soc. America*, Vol. 34, p. 76 (1962).
5. Prestigiacomo, A. J. and MacLean, D. J. "A frequency shifter for improving acoustic feedback stability", *J. Audio Eng. Soc.*, Vol. 10, p. 110 (1962).
6. Bode, H. and Moog, R. "A high-accuracy frequency shifter for professional audio applications", *J. Audio Eng. Soc.*, Vol. 20, p. 453 (1972).
7. Dome, R. B. "Wideband phase shift networks", *Electronics*, Vol. 19, p. 112 (Dec. 1946).
8. Luck, D. G. C. "Properties of some wideband phase-splitting networks", *Proc. I.R.E.*, Vol. 37, p. 147 (Feb. 1949).
9. Orchard, H. J. "Synthesis of wideband two-phase networks", *Wireless Engineer*, Vol. 27, p. 72 (March 1950).
10. Stein, B. B. "Investigation and calculation of a wideband phase-shift network" (in German), *Nachrichtentechnik*, Vol. 4, No. 12, p. 513 (Dec. 1954).

H.F. Predictions for July

The extended periods of magnetic activity which have been observed each month so far this year should now begin to abate. During these periods maximum usable frequencies are reduced and on routes crossing latitudes greater than about forty degrees lowest usable frequencies are increased.

The charts are based on an ionospheric index of 31; this is one third of the value for July of the last year — the effect being to lower HPF and FOT curves one or two megahertz.