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THE IMPLEMENTATION AND PERFORMANCE ENHANCEMENT OF A COMPLETELY DIGITAL POWER AMPLIFIER

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Abstract

The paper addresses the problems associated with producing a purely digital power amplifier which is based upon the class D (Pulse Width Modulation) mode of operation, which is an amplification technique particularly suitable for this application. An enhanced digital pulse width modulation conversion process is introduced and is shown that the use of such a conversion process can lead to a substantial reduction in the harmonic distortion introduced by the sampling technique when compared with existing techniques. The practical digital implementation of the conversion process is also described using a system which consists of a dedicated digital signal processor together with very high speed digital logic circuits.

Introduction

For some years there has been a desire to produce an all digital audio system. To date only the power amplifier and the loudspeaker system halts the progress towards this goal. The digital amplifier based upon the class D mode of operation proposed here is one in which the input is in the form of digitised samples of an audio signal which are then digitally processed in such a way to produce a power digital signal which after filtering can be directly applied to a loudspeaker system. This system alleviates the need for conventional digital to analogue conversion followed by linear analogue amplification. The basic form of the digital class D amplifier is shown in figure (1). The digital audio samples are subject to some form of digital preprocessing and then are converted directly to pulse width modulation. The resultant low voltage pulse train can than be amplified easily and efficiently by a power output switching stage which uses devices such as power MOSFETs, which have a capability for high switching speeds. Finally the amplified signal is restored when the output of the power stage is applied to an analogue lowpass filter.

Class D power amplification itself has a number of distinct advantages over conventional linear power amplifiers primarily due to the method of operation. These include :

1. The switching nature of the power output stage leads to efficiencies upwards of 80 per cent, therefore leading to a smaller power supply and smaller heatsink.

2. The simple repeatable construction and lower complexity of the amplifier design leads to lower manufacturing costs.

3. The amplifier can be made physically smaller than a linear power amplifier of comparable power output especially when designed with a switched mode power supply, and is therefore particularly suitable for P.A applications were large numbers of high power output amplifiers are required .

4. Maintaining digital operation throughout allows the use of a dedicated digital signal processing techniques which could be used to perform additional acoustical functions such as tone control, reverberation etc.

However it is fair to say that the digital class D power amplifier has two major disadvantages.

Firstly the digital circuits involved in the pulse width modulator need to operate at very high speeds which will ultimately limit the fundamental resolution of the system. Secondly extra distortion components are added to the signal when the digitised samples are converted to pulse width modulation. The paper discusses a sampling process which lowers the distortion present in the pulse width modulated signal and shows how it can be implemented using a digital signal processor and high speed ECL logic.

Digital Implementation of Pulse Width Modulation

The use of class D amplification dictates that its input signal must be converted to a pulse width modulated format. This can be achieved with a system of the form shown in figure (2). Digital counters are loaded by a digital preprocessor with a value at fixed intervals separated by some time T, whose frequency determines the carrier frequency of the pulses. The counter would then count down until a terminal count i.e 0 is reached taking a time defined as T_p. At this point a timing pulse, derived from the counter, is used to change the state of a J-K flip flop device which would produce a pulse edge and therefore defining a pulse width given by T_p . Three alternative techniques based on this principle are shown diagrammatically in figures (3a) to (3c). Figure (3a) shows how pulses with trailing edge modulation are derived, which is generally the most simple case. Symmetrical double edge modulation can be formed by a similar method as shown in figure (3b). In this case pulses are produced having edges that are equidistant from the centre timing points where the leading edge is defined by loading the counters with a value equivalent to the time $(T/2 - T_p)$. Another form of modulation, again double edged, is shown in figure (3c) where separate samples are used to define each pulse edge in contrast to the previous case where a single sample defined the position of both edges. In this case the pulse edges are asymmetrically positioned about the centre timing points.

From the above description the task converting digitised samples to pulse width modulation would seem to be relatively simple. However the digital circuits involved in the processing have to run at a very high clock speed. For example in a 16 bit system with a carrier frequency of 44.1 kHz the clock frequency necessary to drive the counters is : $F_{clk} = 2 \times 44.1 \times 10^3 \times 2^{16} = 5.78GHz$ which is clearly not easily practically attainable.

If a resolution of 12 bits would suffice then this figure would be reduced to $361.27MHz$ which is achievable with ECL device technology. It can be seen that it is desirable to keep the pulse width modulation carrier frequency as low as possible in order to keep the clock frequency a minimum. A 12 bit system has been built based around TTL Fast and ECL logic in order to implement these ideas together with a technique for reducing the clock frequency without sacrificing the fundamental resolution of the system. The system is spilt into two sections consisting of a TTL section and an ECL logic section in which the TTL logic is used primarily to produce pulse width modulation of 9 bits resolution. ECL logic is then used to effectively change the position of the pulse edges according the numerical value of the remaining 3 bits. A block diagram of the system is shown in figure (4).

As mentioned earlier resolution to 12 bits would normally require a clock frequency of approximately $360MHz$ however a logic circuit has been devised, shown in figure (5), that can achieve a similar resolution but with a fundamental clock frequency of $90MHz$. The timing waveforms that are applied to this circuit consist of the fundamental clock signal as well as quarter cycle delayed, inverted, and half frequency versions of it, these waveforms are shown in figure (6). Ideally the circuit driven by the 3 least significant bits delays the pulse edges in steps of $2.77ns$ a maximum of $22.16ns$. On test it has produced pulses with a maximum error of $2ns$ which corresponds to a 0.7 bit error although in most cases produces less than a 0.3 bit error.

Sampling Techniques

Production of pulse width modulation is normally categorised into two forms which are entitled *Uniform Sampling* and *Natural Sampling* which are shown in figures (7) and (8) respectively. In figure (7) it can be seen that if uniform sampling is used the widths of the modulated pulses are proportional to the instantaneous amplitude of the modulating signal at uniformly spaced intervals of time. The modulating signal is shown here as a sample and hold waveform were it can be seen that its amplitude does not change throughout the sampling interval. It is important to note that the sampling instants are not coincident in time with the edges of the modulated pulses. The application of uniform sampling is particularly suitable here since its use relies on sampled signals for the production of pulse width modulation. In contrast natural sampling uses analogue signals for the production of pulse width modulation. Pulse width modulation of this form is produced when signals of the form in figure (8) are applied to an analogue comparator. The resulting pulse width modulated signal would have pulse edges that are coincident in time with the sampling instants and that the time between sampling instants can vary. Natural sampling cannot be implemented digitally since complete knowledge of the signal between sampling instants is required. However it can be approximated in the digital domain with the use of linear interpolation.

Both sampling techniques can be used to produce modulation on either or both edges of the pulse. However from this point in the paper the discussion is confined to double edge modulation as it ultimately results in higher performance than single edge modulation and is therefore more applicable for audio use.

Spectra of Modulated Pulse Trains

The use of uniform sampling or natural sampling gives pulse trains that have distinctly different frequency spectra. It can be shown that double edge uniform sampling pulse width modulation has the frequency spectrum given by equation (1):

$$
\begin{aligned}
F(t) = {} & \sum_{n=1}^{\infty} \frac{2HJ_n(Mn\pi\alpha/2)}{(n\pi\alpha)} \sin(n\pi/2)\cos(n\omega_v t - n\pi\alpha/2) \\
& + \sum_{m=1}^{\infty}\sum_{n=-\infty}^{\infty} \frac{2HJ_n(M\pi(m+n\alpha/2))}{\pi(m+n\alpha)} \sin((m+n)\pi/2)\cos(n\omega_v t + m\omega_c t - n\pi\alpha/2)
\end{aligned} \tag{1}
$$

Similarly the frequency spectrum when natural sampling is used is given by equation (2) :

$$
F(t) = \frac{H\cos\omega_v t}{2} + \sum_{m=1}^{\infty}\sum_{n=-\infty}^{\infty} \frac{2HJ_n(M\pi m/2)}{(\pi m)} \sin((m+n)\pi/2)\cos(n\omega_v t + m\omega_c t) \tag{2}
$$

where:
α = ratio of signal to carrier frequency
ω_c = carrier frequency
ω_v = signal frequency
J_n = Bessel function of the first kind with integer order

H = pulse height
M = modulation depth
m = carrier harmonic number
n = signal harmonic number

Referring to equations (1) and (2) the nature of the distortion added to the original signal can be categorised into two forms, defined as follows :

1. Forward harmonic distortion, where the distortion components are harmonically related to the input signal. These components are given by the first terms in the above expressions.

2. Combinational folded back distortion, where the distortion components are at frequencies centred around multiples of the carrier frequency and therefore are non-harmonically related to the input signal. These components are given the second terms of (1) and (2).

The distortion components that are of interest are those which fall within the passband of the amplifier, itself a function of the bandwidth of the output lowpass filter. Typically these would be forward harmonic components with frequencies given by $n\omega_s \le \omega_{bp}$ and combinational components with frequencies given by $(\omega_c - n\omega_s) \le \omega_{bp}$, where ω_{bp} is the break point of the output lowpass filter. In practice the finite rate of roll off of the filter also needs to be accounted for.

Referring to equations (1) and (2), a distinct difference in the frequency spectrum of the modulated pulses of both natural sampling and uniform sampling can be seen. Natural sampling leads to zero amplitude forward harmonics with all the distortion coming from the combinational components. In contrast the use of uniform sampling results in the production of both forward harmonic and combinational distortion. However the amplitude of the combinational components are lower than those obtained when using natural sampling. A quantitative description is shown in figure (9) which represents an amplifier with a carrier frequency of F_c subject to sinusoidal input of frequency F_1 . As can be seen there is ensemble of distortion components which fall within the envelope of the lowpass filter and therefore are not attenuated. The use of asymmetric uniform sampling produces only odd order

combinational components, $(F_c - nF_1)$ n odd, and also for symmetric uniform sampling the addition of even order forward harmonic components aswell. The uniform and natural sampling processes discussed above result in a different combination of forward harmonic distortion and combinational folded back distortion and so the choice of sampling technique is particularly important. There are various factors which influence this choice some of which are:
1:Bandwidth of the input signal.
2:Modulating (carrier) frequency.
3:Modulation depth and the maximum signal strength
4:Amplitude response of the output lowpass filter.

The digital class D power amplifier proposed would ideally have a carrier frequency of 44.1 kHz or 88.2 kHz (an integer multiple of the sampling frequency of compact disc systems). Any higher would introduce extra problems associated with the speed of the digital hardware and also extra energy losses in the power output stage. Studies have shown that at these intermediate carrier frequencies neither uniform or natural sampling result in the best performance. This has led to the consideration of an enhanced sampling technique, similar to that suggested by Malanov [2] and presented in [3], that is controllable which can take a range of forms between uniform sampling and natural sampling. This process, depicted in figure (10), may be described by considering a signal f(t) whose sampled values at times nT and $(n+1)T$ are S_1 and S_2 respectively. The pulse width at time nT is derived by the natural sampling of a transformed version, $f_n(t)$, of the original signal $f(t)$ between these samples , defined as follows:

$$f_n(t) = f((1-\varepsilon)(t-nT)+nT) \quad 0 \le \varepsilon \le 1 \tag{3}$$

Here the variable ε relates to the extent of departure of this process from natural sampling. When ε is zero natural sampling is obtained, alternatively setting ε to unity achieves uniform sampling. Although it be would difficult to implement this process exactly in practice, it is considered here since an analytical expression can be obtained for the spectrum of the PWM it produces and it is illustrative of the sampling process in which the distortion terms can be varied, in this case by the factor ε. The spectrum for the new sampling process can be written in general form as:

$$F(t) = \sum_{n=1}^{\infty} \frac{2HJ_n(Mn\pi\alpha\varepsilon/2)}{(n\pi\alpha\varepsilon)} \sin(n\pi/2)\cos(n\omega_s t - n\pi\alpha\varepsilon/2)$$
$$+ \sum_{m=1}^{\infty}\sum_{n=-\infty}^{\infty} \frac{2HJ_n(M\pi(m+n\alpha\varepsilon/2))}{\pi(m+n\alpha\varepsilon)} \sin((m+n)\pi/2)\cos(n(\omega_s t) + m\omega_c t - n\pi\alpha\varepsilon/2) \quad (4)$$

It follows that setting $\varepsilon = 1$ forces the above expression to that of equation (1) and similarly setting $\varepsilon = 0$ reduces it to the form of equation (2). With intermediate values of ε there is a tendency for the forward harmonic distortion terms to increase with ε and the combinational terms to reduce with ε.

Implementation of Enhanced Sampling

An approximate form of enhanced sampling is obtained when the digitised input samples are applied to the transformations given by equations (5) and (6). The timing of the trailing edge of the pulse is derived from the normalised samples S_1 and S_2 of the signal and the sampling period T, shown in figure (7), defined as T_{p1} is given by :

$$T_{p1} = \frac{S_1 + 1}{2 - (1-\varepsilon)(S_2 - S_1)}\, T/2 \quad (5)$$

Similarly for the leading edge, T_{p2} is given in terms of the samples S_2, S_3

$$T_{p2} = \frac{1 - S_2}{2 - (1-\varepsilon)(S_2 - S_3)}\, T/2 \quad (6)$$

The values of T_{p1} and T_{p2} loaded into the digital counters which generate the pulse width modulation are calculated from the sampled signal values in the digital preprocessor shown in figure (2). A single chip Texas TMS32010E15 device was used for this purpose in the prototype system. Since equations (5) and (6) involve mathematical divisions They are implemented in the TMS software by a relatively simple iteration procedure. This procedure which does not require fixed point division and can be readily implemented in real time by the digital signal processor. The procedure, derived from equations (5) and (6) produces sequences T_{p1}^k and T_{p2}^k for the k^{th} iteration in the following manner and can be accomplished using integer arithmetic.

$$T_{p1}^{(k+1)} = \frac{S_1 + 1}{2} + ((1-\varepsilon)\frac{(S_2 - S_1)}{2})T_{p1}^k \quad (7)$$

and

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$$T_{p2}^{(k+1)} = \frac{1 - S_2}{2} + ((1-\varepsilon)\frac{(S_2 - S_1)}{2})T_{p2}^{k} \qquad (8)$$

The above equations converge, to the required values of T_{p1} and T_{p2} respectively after approximately five iterations from initial values $T_{p1}^0 = S_1$ and $T_{p2}^0 = S_2$.

Results

Previous work [4] has indicated that with apt design the power output stage of the class D amplifier need not introduce any significant distortion components to the modulating signal. Therefore distortion that is produced can be seen to arise mainly from the pulse width modulation conversion process itself. The results presented here are indicative of the distortion produced by the conversion process.

The system under test has a carrier frequency of 44.1 *kHz* and is subject to sinusoidal modulation with frequencies up to 20 *kHz*. Output filtering is accomplished by a 6^{th} order Butterworth lowpass filter with a 3dB point at 20 *kHz* In order to assess the maximum distortion that would be produced the amplitude of the sinusoids were chosen so as to set the modulation index equal to 0.9 which is similar to a full power output test of a conventional linear power amplifier. Figures (12) and (13) show the level of total harmonic distortion produced as a function of the sampling factor ε when sinusoids with frequencies of 4, 6, 10 and 20 *kHz* are used as the modulating signals. Figure (12) is obtained for the ideal sampling technique described by equation (4) whereas figure (13) presents the distortion measured from the practical system. The level of distortion for 10 kHz and 20 kHz inputs appears to be extremely high however it must be noted that in typical audio music signals the power as a function of frequency decays at approximately 6 decibels per octave. This results in a reduction of the modulation index at these higher frequencies which in turn reduces the level of distortion. The modulation index at 20 kHz would normally be only 0.1 or less and typically 0.2 at 10 kHz resulting in distortion levels some 20 dB less than those shown in the figures. The advantage of the enhanced sampling technique over conventional uniform sampling can be seen as a reduction of the level distortion by up to 12dB at the lower frequencies. If the reduction of modulation index with increasing frequency is also taken into account then an optimum value of ε is seen to emerge. Here a value of $\varepsilon = 0.6$ would result in the best across the band performance. It can be seen that there is good agreement between the ideal and practical implementation for the 10 *kHz* and 20 *kHz* results for all values of ε and similarly for the 4 *kHz* and 6 *kHz* results from $\varepsilon = 1$ down to 0.4. Below this value the experimental results deviate from those expected which has been identified as an additional third harmonic distortion component as a result of the linear interpolation approximation.

Conclusion

The digital implementation of a class D power amplifier has been considered together with a qualitative description of the distortion such an amplifier would produce. An enhanced sampling process has been shown which can lead to an overall reduction of distortion compared with more conventional sampling techniques. The digital implementation of this technique, involving the use of a digital signal processor, has been discussed. The use of the enhanced sampling technique has resulted in an acceptable performance being obtained from the full bandwidth system which has a carrier frequency of 44.1 *kHz*. This type of digital amplification is seen to have applications in P.A and in-car audio systems.

References

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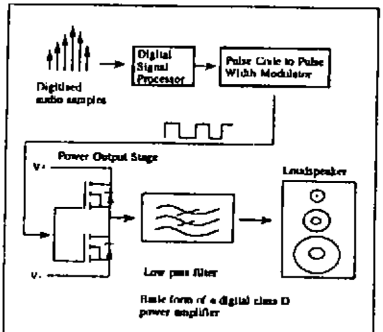
[2]. Mananov, V.V.: 'Single polarity pulse width modulation of the third kind', Telecomm. Radio Engineering, Vol 22, Pt 2, 1967, pp. 67-70
[3]. Leigh, S.P., Mellor, P.H., Cheetham, B.M.G.: 'Distortion minimisation in pulse width modulated systems using a digital sampling process', Electronics Lett., 1990, 16, p. 1310
[4]. Attwood, B.E.: 'Design parameters important for the optimisation of very high fidelity PWM (Class D) audio amplifiers', Journal of the Audio Engineering Society., 1983, 31 no. 11, pp. 842-853



Basic form of a digital class D power amplifier

Figure (1)

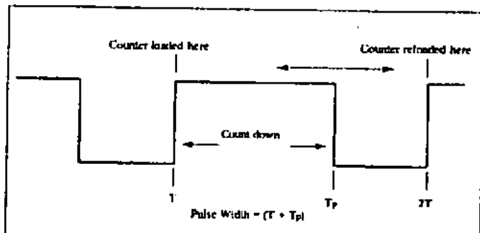


Figure (3a) : Trailing edge modulation

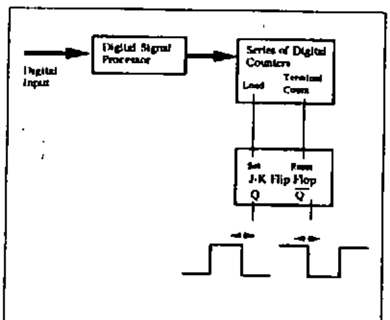


Figure (2)

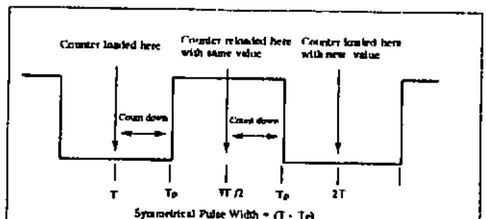


Figure (3b) : Symmetrical double edge modulation

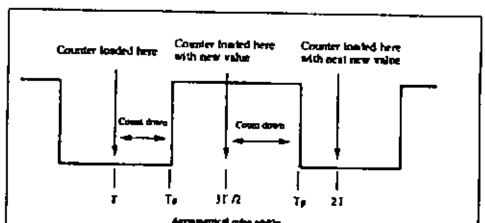


Figure (3c) : Asymmetrical double edge modulation

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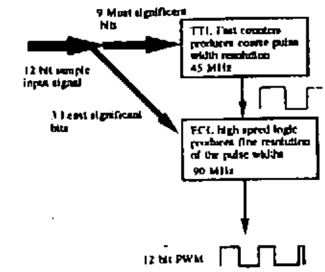


Figure (4)

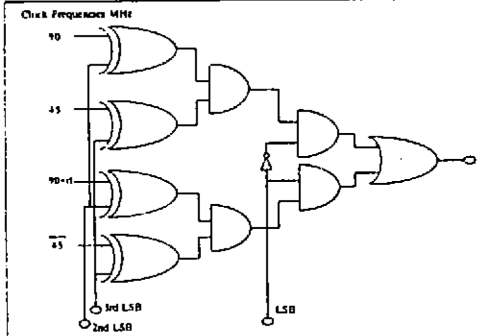
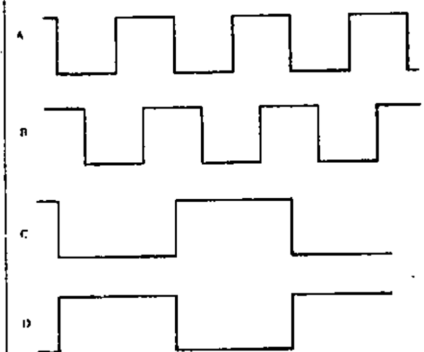


Figure (5) : Pulse resolution circuit



A 90 MHz clock signal
B 90 MHz + 1/4 cycle delay
C : 45 MHz
D : 45 MHZ Inverted

Figure (6) : Timing waveforms

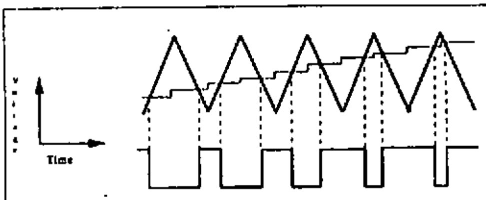


Figure (7) : Uniform Sampling

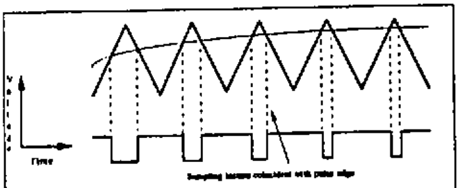


Figure (8) : Natural Sampling

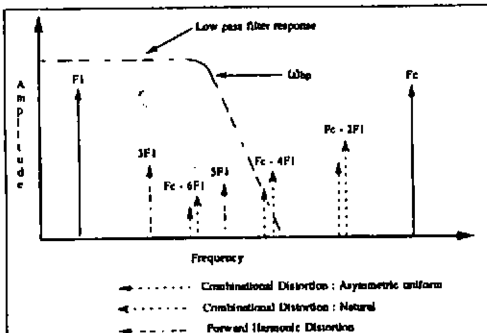


Figure (9) : Distortion products

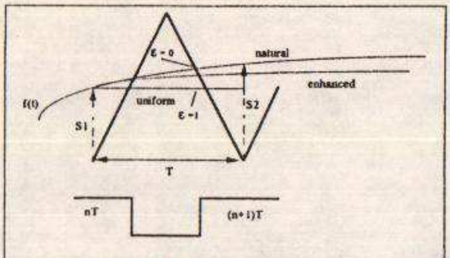


Figure (10) : Enhanced sampling

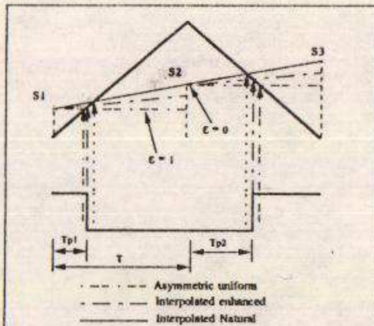


Figure (11) : Linear interpolation of enhanced sampling

Fig (12): Theoretical results: Carrier frequency = 44 kHz, Butterworth 6th order lowpass output filter Fc = 20 kHz, M = 0.9

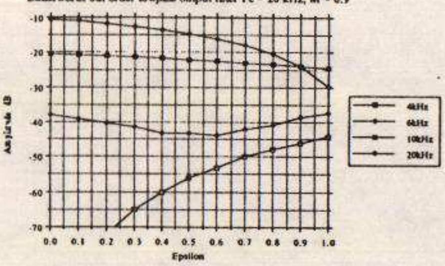


Fig (13): Experimental results ; Carrier frequency = 44 kHz, Butterworth 6th order lowpass output filter Fc = 20 kHz, M = 0.9