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ELECTRONICS WORLD

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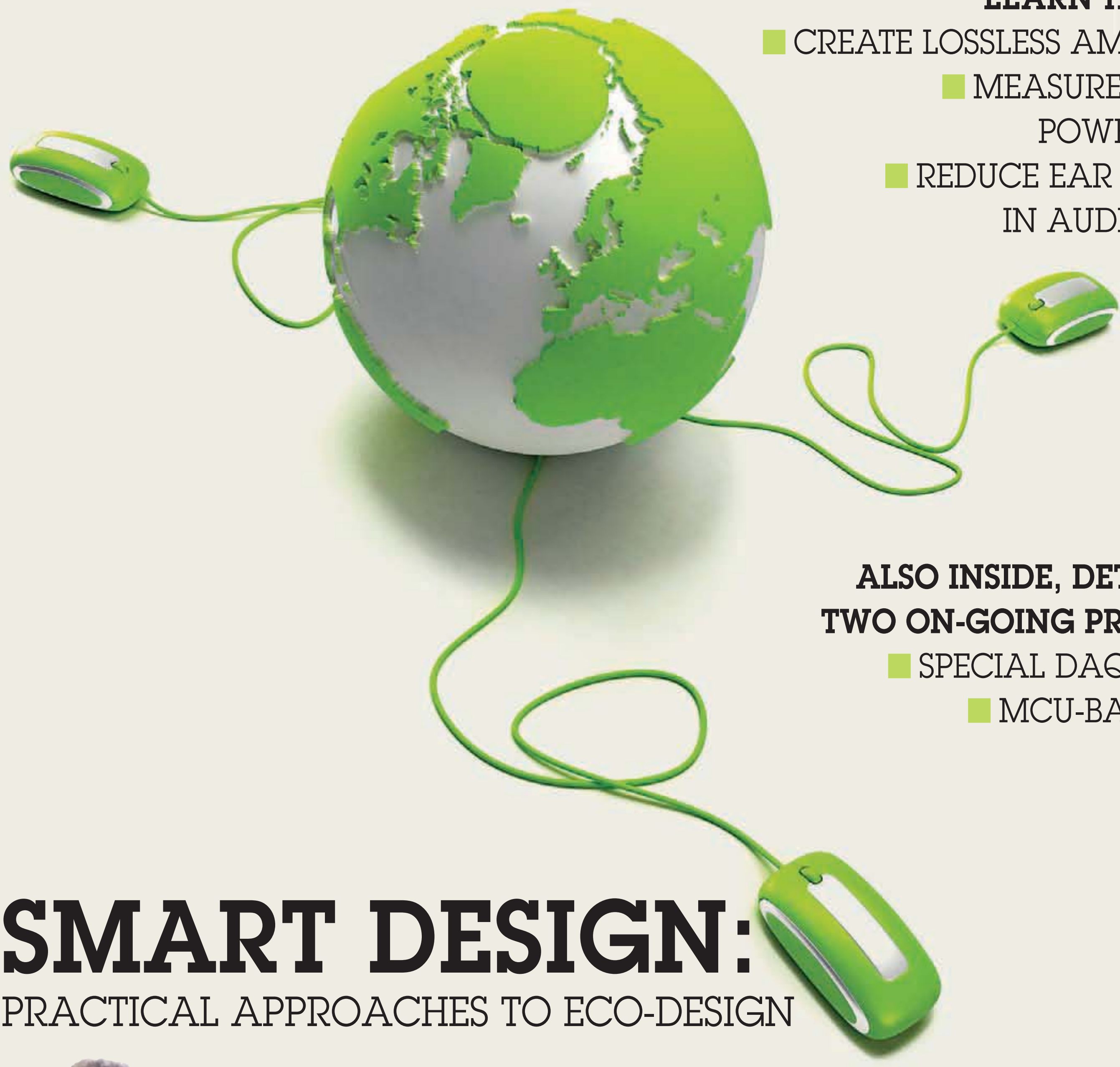
THE ESSENTIAL ELECTRONICS ENGINEERING MAGAZINE

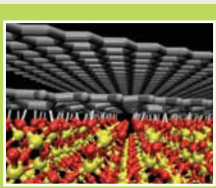


IN THIS ISSUE
LEARN HOW TO:
- CREATE LOSSLESS AMPLIFIERS
- MEASURE HUM IN POWER AMPS
- REDUCE EAR FATIGUE IN AUDIO AMPS

ALSO INSIDE, DETAILS OF TWO ON-GOING PROJECTS:
- SPECIAL DAQ BOARD
- MCU-BASED PLC

SMART DESIGN:
PRACTICAL APPROACHES TO ECO-DESIGN

MYK SAYS
TEST YOUR RECEIVER

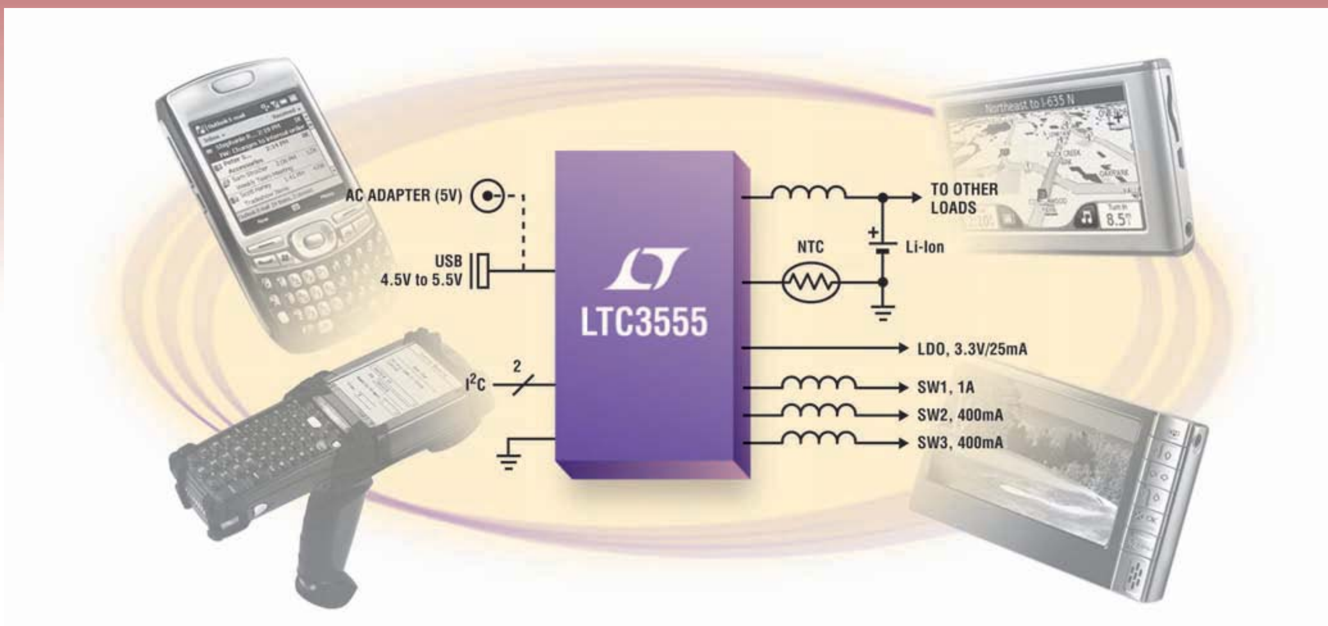


TECHNOLOGY
GRAPHENE'S DOUBLE NATURE CONFIRMED BY RESEARCHERS

PRODUCTS
DIGITAL MULTIMETERS, LED DRIVER ICS, NANO CONNECTORS

ALSO IN THIS ISSUE: THE NEW COLUMN FROM THE NANOTECHNOLOGY KNOWLEDGE TRANSFER NETWORK

What Portable Power Problem?



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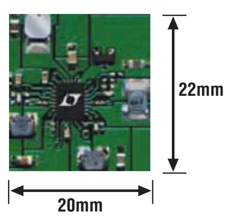
LTC3555 Demo Circuit



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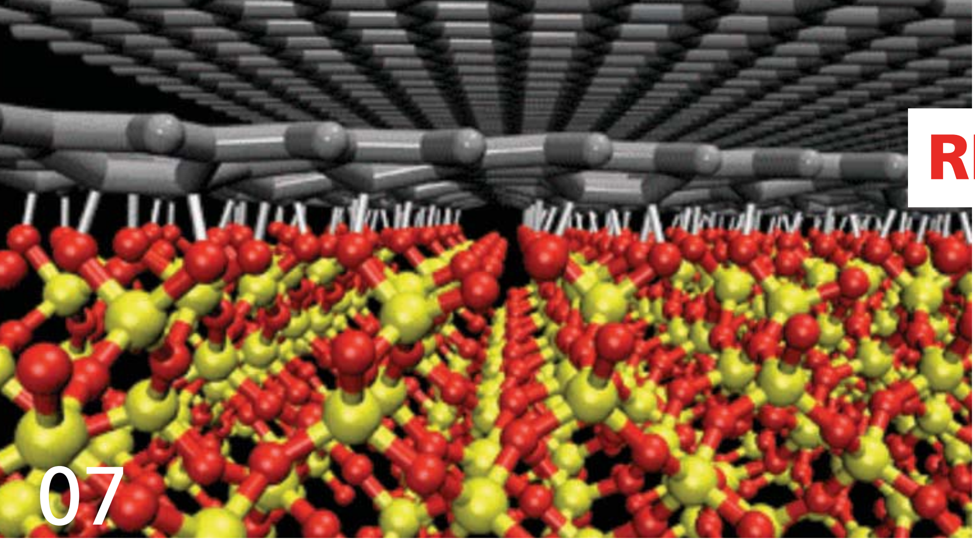

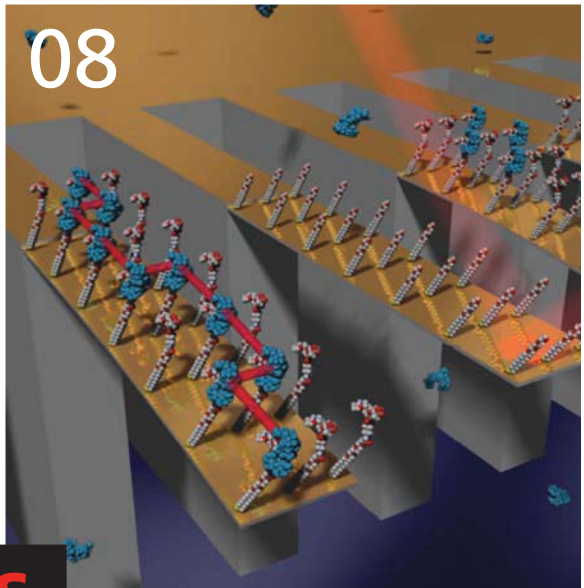

REGULARS

FEATURES

Find about Environwise's smart design webinar on **Page 11**

ELECTRONICS WORLD

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Carry on Doing What You Do Best

Dear Readers,

Recently we (journalists) were invited to a press briefing by a young company to be introduced to a new type of power-saving technology for handheld devices. The technology has already been successfully used in communications infrastructure equipment, but seeing the power and costs savings the technology can bring to the systems it is installed in, the developers decided to tap into the lucrative market of handheld systems too, where power and cost are of crucial importance.

However, the journalists' reaction was not as expected; they questioned the success of a new technology in an overcrowded market where established, well-known names tend to dominate. This could potentially reflect the feelings of OEMs too.

So the question is: Is there room for new devices and IP in markets that are already overcrowded and dominated by a handful of well-known companies?

The answer should be 'yes'. There should always be a place for new technology, regardless of the foothold big names already have in that market.

Where would many companies be now if they were giving up in their earlier days? To a degree a rebellious nature is as important in any technology developer as an entrepreneurial spirit is. And, as any engineer will tell you, the excitement felt when something new is developed is overwhelming and will not go unnoticed by other engineers and engineering firms.

Equally, however, it needs to be emphasised that the marketing of any new technology is key – actually, a very powerful marketing machine is what is needed to make something succeed in an already overcrowded market place.

All I'd like to say is 'Good Luck' and 'Never Give up'!

Editor
Svetlana Josifovska

OVER THE PAST SEVERAL MONTHS WE HAVE BEEN RUNNING A READER OFFER BY **ANDERS ELECTRONICS** IN THE MAGAZINE AND ON LINE.

That competition is now closed and the winners selected, which are:

Nigel Lee, Engineering Manager from MacGregor Welding System Ltd, UK, winner of the Anders Electronics's EVK-UMRX10-B-70 with Wi-Fi

AND

Peter Wortley, Senior Technologist at Triton Technology Ltd, UK. Peter is the winner of Anders Electronics's EVK-UMRX10-D-70 with Wi-Fi.

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787

■ Telemedicine is gathering momentum in Germany, where doctors are increasingly checking on pacemakers via the Internet and give advice on emergency treatments via video links.

The term "telemedicine" refers to medical services that are delivered at a distance with the aid of modern information and communication technology (ICT). This kind of remote care was only possible under experimental conditions, especially as it raised difficult legal and funding issues.

But since last year, doctors in Germany are being paid for the remote monitoring of patents with medical devices, through a billing tariff; billing codes given to GPs allow them not only to use telemedicine in certain situations, but also to charge for it.

■ Vodafone has successfully trialled an HSPA + 64QAM mobile broadband technology at peak data download rates of 16Mbps.

The company now plans to trial mobile broadband data connections with peak rates of up to 21Mbps early this year using HSPA + MIMO functionality. Its experts estimate that the technology would be capable of delivering a typical video download experience of more than 13Mbps in good conditions and an average of more than 4Mbps across a full range of typical cell locations including urban environments. If the trials prove a success, Vodafone plans to make this technology available in selected commercial networks.

■ The Californian government, Intel and Hewlett Packard are jointly setting up a symposium on the global prospects and trends in 'greener' electronics supply chains for 2009.

The symposium, entitled "*The Greening of Electronics in a Global Economy*" will be held on February 19, 2009, at the California Environmental Protection Agency's headquarters in Sacramento. It will address the need and opportunities for reducing hazardous substances in consumer electronics, current US and international best practises in health and environmental compliance practices, measures open to electronics manufacturers for ensuring compliance and global prospects and trends in 'greener' electronics supply chains and new regulation.

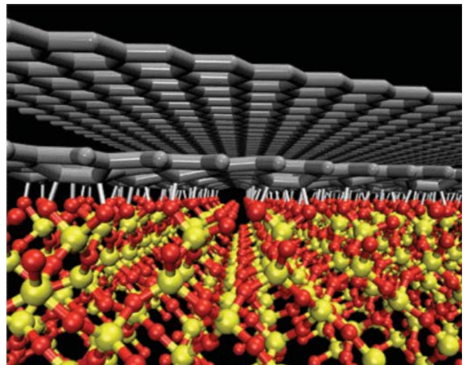
GRAPHENE CAN BEHAVE WITH METALLIC OR SEMICONDUCTOR PROPERTIES, DISCOVER RESEARCHERS

Researchers at the Rensselaer Polytechnic Institute have discovered a new method for controlling the nature of graphene, bringing academia and industry one step closer to realizing the mass production of graphene-based nanoelectronics.

Large-scale quantum mechanical simulations have shown that when deposited on a material treated with hydrogen, graphene exhibits metallic properties; whilst when deposited on a surface treated with oxygen, graphene exhibits semiconductor properties, indicating that the chemistry of the surface on which graphene is deposited plays a key role in shaping the material's conductive properties.

"Depending on the chemistry of the surface, we can control the nature of the graphene to be metallic or semiconductor. Essentially, we are 'tuning' the electrical properties of the material to suit our needs," said Saroj Nayak, an associate professor in Rensselaer's Department of Physics, who led the research team.

The new method for "tuning" the nature of graphene is a key step to separating the graphene's properties, as conventionally whenever a batch of graphene nanostructures is produced some of it is metallic, while the rest is semiconductor. To date it has been nearly impossible to separate the two on a large scale, even though what's required is that the graphene devices be comprised solely of metallic or semiconductor graphene.



The rendering of two sheets of graphene, each with the thickness of just a single carbon atom, resting on top of a silicon dioxide substrate

Graphene is a one-atom-thick sheet of carbon, discovered in 2004. The material's conductive properties make it attractive to researchers as even at room temperature, electrons pass effortlessly near the speed of light and with little resistance. This means a graphene interconnect would likely stay much cooler than a copper interconnect of the same size in ICs and electronic systems.

Shoppers' Lament for the Old Incandescent Bulb

There's been a mad shopping rush on conventional type of incandescent light bulbs in the UK, as the main supermarket chains decide to voluntarily stop selling them ahead of the EU's planned phase-out in September.

The mad rush in which people have been bulk-buying the bulbs is explained by the fact that the new type of energy-saving bulbs do not fit inside many of the old fittings, and some of the new type of bulbs' lifetime proves much shorter than advertised.

According to UK's Department of Environment Food and Rural Affairs (Defra) however, only some 5% of the all sockets are incompatible with the new energy-saving lights, even though only a couple of years ago it reported that figure to be much higher at 50%.

The switch to energy-saving type lighting is hoped to reduce carbon dioxide emissions by nearly five million tonnes a year, making the emissions cut close to 30% within 10 years.

The first to go were the 150W bulbs in 2007, followed by the 75W bulb and the 60W bulb this year.

The fluorescent lights, which will be the only type of bulb available by 2012, are also incompatible with older ceiling sockets. They do not work with most dimmer-switches and security timers, are not suitable for chandeliers and can't cope with freezing conditions when used outdoors.

The low energy alternatives are compact fluorescent lights (CFLs), like small versions of the fluorescent strip lights found in offices and public buildings, and halogen lights shaped like ordinary bulbs. They use just a fifth of the energy of a conventional bulb.

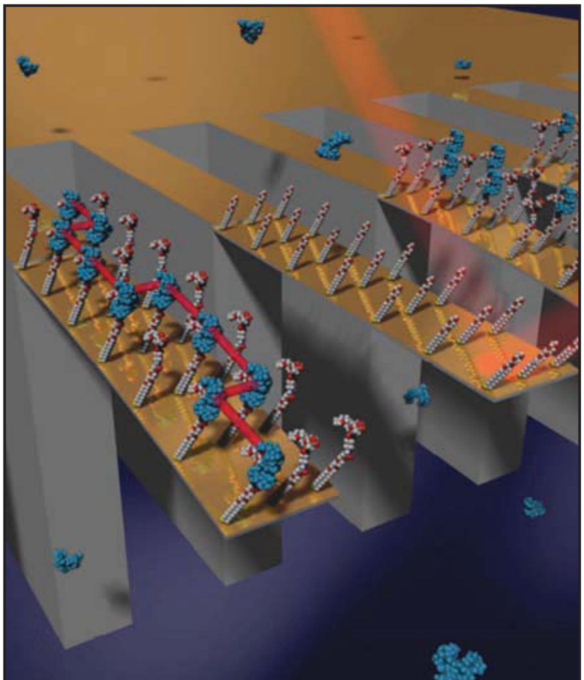
EXPLORING THE INTERSECTION BETWEEN AND THE LIFE SCIENCES

BY MIKE FISHER

The Nanotechnology Knowledge Transfer Network (NanoKTN), one of the UK's primary knowledge-based networks for Micro and Nano-technologies, was set up by the Technology Strategy Board to promote and facilitate knowledge exchange, support the growth of UK capabilities, raise awareness of nanotechnology and provide thought leadership and input to the UK policy and strategy.

The NanoKTN's activities are built around focus groups which identify the gaps in the supply chain, as well as identifying the UK's potential in innovation. This information is reported back to the Technology Strategy Board to input into their UK Nanotechnology Strategy and also provides leverage for channeling government funds into specific areas of need.

For further information on the NanoKTN and its activities visit www.nanoktn.com



Micro cantilevers for drug binding measurement
[Courtesy of the London Centre for Nanotechnology]

ON THE 11TH of February, the Nanotechnology Knowledge Transfer Network (NanoKTN) and The Wellcome Trust are bringing together members of the nanotechnology and life science communities at the Nano4Life conference to explore the convergence of these two fields.

Micro and Nanotechnologies (MNT) are increasingly entering the healthcare field, as significant research funding has been provided by governments over the past ten years in the race to capture the potential intellectual and economic benefits that MNT has been promising. These research efforts have produced numerous advances that are starting to be adopted by the healthcare industry. The diagnostics field in particular has benefitted from this miniaturisation trend.

There are two major issues faced by western healthcare markets that these technologies aim to address. The first is caring for an ageing population who wish to remain active and healthy. This is driving up demand within healthcare systems. The second is the desire of governments and payers to curb spending within their healthcare budgets as greater regulatory hurdles increase the cost of getting products to market. This is placing pressure on healthcare providers to find efficiencies and cut costs wherever possible. Miniaturisation is helping solve these issues, as products become more specific; use fewer reagents but also lower their use as they are expensive.

The development of microfluidics has allowed the use of volumes thousands of times smaller than a common droplet. This technology forms the basis of lab-on-a-chip devices for sensitive analytical measurements. The reduced volumes used by lab-on-a-chip systems not only enables less reagents and samples to be used, but also decrease analysis times, as actions like heating, cooling and mixing are significantly faster than in traditional macro systems. In addition, smaller devices allow the production of cheap disposable systems, which prevent cross contamination and increase accuracy.

Lab-on-a-chip devices have received significant attention in the research community in recent years, and this has produced a large amount of intellectual property and numerous spin-out companies. One of the new areas being considered is that of moving into nanofluidics, where the surface tensions and adsorption issues are extremely challenging.

The ultimate aim is to detect diseases at the cellular level at the early stages, before they require costly treatments. An example to be showcased at the Nano4Life conference is the use of nano and micro technology by Sphere Medical to produce micro analysers in closed loop systems within the critical care setting. These aim to decrease mortality in the critical care setting by closely monitoring patients and enabling rapid, automatic responses to changes in clinical measurements.

Another area of diagnostics to be discussed during Nano4Life is being researched at the University of Glasgow. Here

NANOTECHNOLOGY

Professor Jon Cooper is looking at systems which can be integrated into a ‘lab-on-a-pill’. This device is being designed to be swallowed by the patient to provide diagnostic information as it travels through the digestive tract. Such systems require the development of new miniaturised power sources, imaging and detection systems. Another major consideration is how to reliably transmit and receive communications from such low-power devices, especially from implanted and internalised devices, as low power signals do not pass through tissue easily.

MNT is also being adopted in the development of pharmaceuticals and biotechnology products. Use of miniaturised assay systems, similar to those used for diagnostics, are enabling greater throughput in screening, decreasing the use of expensive reagents and speeding the discovery process.

Work by Dr Rachel McKendry’s group at the London Centre for Nanotechnology has shown that the use of microcantilevers for measuring drug binding can provide greater details on the mechanism of action for antibiotics. In addition, the use of nano-scale particles in the formulation of drugs has decreased toxicity and increased the efficacy of drugs. Both UK-based large pharmaceutical companies, AstraZeneca and GlaxoSmithKline, will be presenting their views of nanotechnology at Nano4Life.

The question of which specific technology will be adopted and, with regards to diagnostics which of these systems will produce the information clinicians and patients truly need and, therefore, which become successful and make it into the market, is yet to be answered. One thing is for sure, micro and nanotechnologies are needed to help solve the issues healthcare systems face around the world. The Nano4Life conference is going to explore these opportunities and how companies and entrepreneurs can become part of the solutions doctors, payers and patients are looking for.

Mike Fisher is the Theme Manager for Bionano and Nanomedicine at the NanoKTN. The Nanotechnology Knowledge Transfer Network (NanoKTN) is one of the UK’s primary knowledge-based networks for Micro and Nanotechnologies. **www.nanoktn.com**

envirowise SUSTAINABLE PRACTICES SUSTAINABLE PROFITS

COULD YOU PROFIT FROM SMARTER DESIGN?

To find out, why not join Envirowise's eco-design team, on-line, on Wednesday 25th February at 10am, for their free 'Practical Approaches to eco-design' webinar. To register, visit www.envirowise.gov.uk/webinar.

Eco-friendly packaging is not just good for a business' green credentials, but can also save money.

From phasing out overpackaging, to sourcing locally, brands large and small are making commitments to improving their environmental practices. However, many businesses still labour under the misapprehension that what's good for the environment will be a cost they have to bear.

However, this needn't be the case. Sustainable business experts Envirowise are urging companies to find out how tackling the issue of over-packaging can save them money by signing up for a free online webinar.

Taking place on 25 Feburary, the webinar is mainly aimed at product and packaging designers to help them deal with rising material costs and increasing pressure to improve environmental credentials.

Envirowise eco-design specialist Jenni Rosser says: "It is estimated that some 80 per cent of the cost of a product and its packaging is set at the design stage, so the benefits of thinking about resource efficiency from the outset can be a win-win situation for both the environment and the bottom line."

Webinar participants can take part in a virtual seminar covering practical elements of eco-design including re-design exercises, comments from key stakeholders in the design industry and a live re-design activity decided on the day by participants.

Mark Tosey, director at design agency Bright Green, took part in the last Envirowise eco-design webinar in October 2008. He said: "The webinar was a very useful and engaging tool to communicate some of the issues around sustainable design. All too often seminars tend to deliver in very general terms, whereas the webinar provided visual stimulus plus live discussion and debate. I 'm looking forward to the next webinar to see how the issues can be explored further."

What is eco-design?

Eco-design is an approach which aims to minimise the costs and environmental impact of a product or packaging over its entire life-cycle. Improving resource productivity allows for the production of more goods and services with fewer materials and input from utilities. With reduced amounts of waste, this creates sustainable benefits for the environment, and ultimately reduces business costs.

With the introduction of the EuP (Energy Using Products) Directive and historic regulations such as WEEE, RoHS, the Packaging Waste Regulations and the Essential Requirements, it's becoming increasingly important to consider the design and lifecycle impacts of products and packaging.

Reaping the benefits

The benefits of taking a measured approach to eco-friendly design can include lighter and more easily packed goods, which equates to lower shipping and transportation costs.

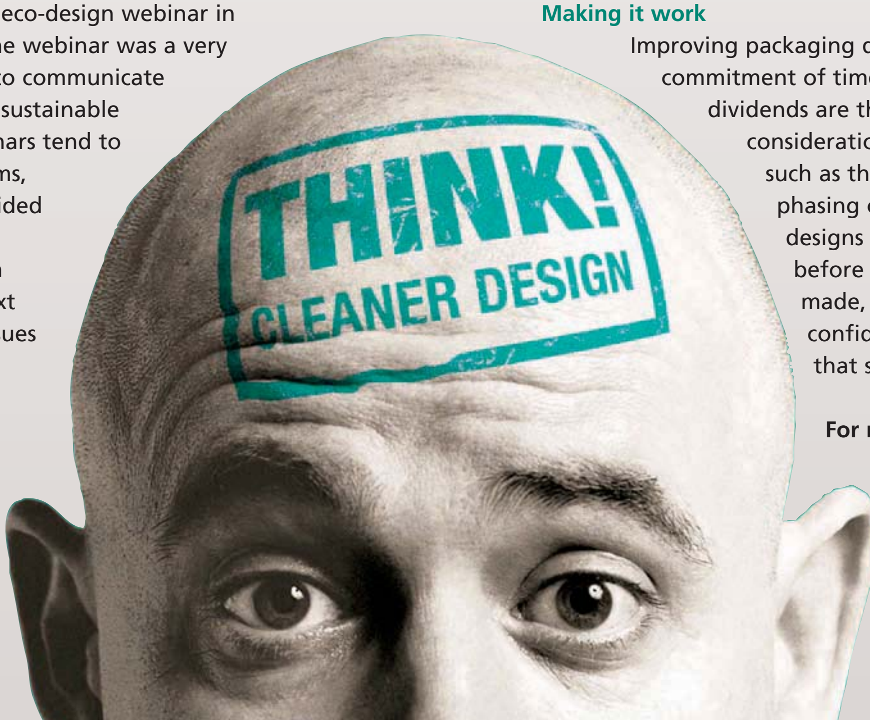
By considering the environment and adopting eco-design techniques, there are opportunities for electrical product designers and manufacturers in particular to play a key role in helping the UK to meet future carbon targets and reduce the emissions associated with inefficient consumer goods.

Designers have a key role to play in helping companies to adopt best practice techniques which could help save around £100m through reduced manufacturing costs for example . They can also assist in gaining a competitive advantage from functionality and service innovation, and can reap the benefits from improved credentials.

Making it work

Improving packaging does require a commitment of time and resources, but the dividends are there to see. By taking into consideration a number of factors, such as the cost of retooling, phasing out existing packaging designs and customer reaction, before wholesale changes are made, a measured and confident plan can be produced that saves money.

For more information on the Envirowise eco-design webinar or to book your place, visit: www.envirowise.gov.uk/webinar or call the Envirowise Advice Line: 0800 585 794



Test your RECEIVER

Myk Dormer

USERS OF UNLICENSED wireless modules are likely to be familiar with the EN300-220 specification, governing the use of such devices on the European ISM bands. Those who have looked a little closer – or who work in the alarm or security industries – will furthermore be aware of the receiver performance classifications detailed in this spec.

To quote the latest release, there are three categories (see section 4.1.1):

"Class1 Highly reliable SRD communication media; e.g. serving human life inherent (may result in physical risk to a person).
Class 2 Medium reliable SRD communication media; e.g. causing inconvenience to persons, which cannot simply be overcome by other means.
Class 3 Standard reliable SRD communication media; e.g. inconvenience to persons, which can simply be overcome by other means, e.g. manual."

According to 300-220, the lowest class of receiver (Class 3) only has to achieve a certain minimum level of RF sensitivity. Class 2 receivers are also required to meet a certain blocking performance level, while the highest performance Class 3 parts must meet a more stringent blocking level and an adjacent channel rejection test also, which corresponds to a receiver performance not much inferior to long-ranged PMR or land mobile radios, governed by EN300-086.

Which is all very well written in the spec, but how do you relate this to the performance of your receiver? How should these parameters be tested and, most particularly, is the receiver you've paid for actually compliant with the class claimed by its manufacturer?

The receiver test methods stipulated in 300-220 are fairly straightforward, although for 'old school' RF engineers used to measuring absolute sensitivity and then relating rejection specs to that, there is a surprise: EN300-220 specifies in terms of absolute interferer levels, not relative values:

"THE RECEIVER TEST METHODS STIPULATED IN 300-220 ARE FAIRLY STRAIGHTFORWARD, ALTHOUGH FOR 'OLD SCHOOL' RF ENGINEERS USED TO MEASURING ABSOLUTE SENSITIVITY AND THEN RELATING REJECTION SPECS TO THAT, THERE IS A SURPRISE"

Sensitivity: A level of -107dBm must yield 20dB or better sinad when measured through a psophometric filter as detailed in ITU-T rec. 0-41.

For data-only links an 0.1% error rate, or 80% message decoding success, is given as equivalent to the 20dB sinad point.

The levels above relate to a 25kHz channel spacing (16kHz BW) radio; for other receiver bandwidths, the limit sensitivity level is given as 10 x log (BW/16) - 107 (dBm).

Rejection: The basic method for all adjacent channel and blocking tests is the same. Remember that the wanted and interfering signal levels are at the combiner output/radio input, not the level at the signal generator.

Two generators are used, feeding the test radio through a combiner or coupler. A wanted signal is applied at a level 3dB higher than the limit sensitivity (-104dBm for a 25kHz unit).

An unwanted signal (unmodulated) is then also applied, at a particular offset and level depending on the test; the measured sinad must remain 20dB or better.

Adjacent channel: This test applies only to Class 1 radios. The interferer has a level of -50dBm and tests are made with this signal one channel above and one channel below the wanted carrier.

Adjacent, saturation: The adjacent channel tests are repeated with the wanted level increased by 43dB (-64dBm for 25kHz) and the interferer at -20dBm. For channel spacings of > 25KHz this interferer has a level of -44dBm in the first test and -10dBm in the second.

Blocking: For Class 1 radios, the interferer has a level of -20dBm and tests are made with this signal 2MHz above and below the wanted carrier. A saturation test is then made, with the same level of interferer but the wanted signal increased by 40dB.

Class 2 radios are tested with a -69dBm interferer at ±2MHz and with a -44dBm interferer at ±10MHz.

Spurious radiation: No receiver is permitted to emit any spuri with levels exceeding 2nW between 9kHz and 1GHz, or 20nW above 1GHz.

Pitfalls: The methods used are fairly straightforward, but there are a few problems in these test techniques that aren't immediately obvious:

1. The audio filter. A psophemetric filter of this type provides a noticeable improvement in sinad compared to the flat 0.3-3.4kHz 'speech bandwidth' filter provided by many sinad-measuring instruments, and a direct

comparison is difficult to make. Ensure you're using the right filter, or you could be throwing away valuable performance margin or failing your radio unfairly.

2. The combiner. Use a good, broadband part. To keep generator output power levels down, use a low loss directional combiner (a part such as the Mini-Circuits ZSC-2-4) with loss around 4dB in preference to a 12dB resistive type. Remember to check your actual signal levels at the combiner output, to allow for cable and coupler losses and other imperfections.
3. Signal generator noise. This is the biggest cause of test error and wasted time in measurement regimes such as this.

The wanted signal generator is uncritical. It is a modulated low level, on-channel signal. Almost any RF source or communication test set will be good enough.

The interfering signal generator is another matter. This requires a good adjacent channel noise level or the measurements will be dominated by the noise sidebands of this generator directly impinging on the carrier. A noise level of around -115dBc/Hz at 25kHz offset or better is required.

When the blocking measurements are made, a different problem becomes evident, as at the high output levels required many signal generators exhibit excessively high wideband noise output. Even middle-range generators can be inadequate.

To check for this generator noise effect, conduct a blocking test on a known, good, high-performance radio unit. Increase the interfering signal level from about 25dB below its eventual level in 1dB steps. A sudden degradation in receiver sinad, out of proportion to the signal level change, indicates the generator noise floor has risen. This will usually be co-incident with an internal attenuator switch over in the generator.

Inexpensive signal generators and communications test set RF sources are rarely adequate as off channel, interfering sources. Middle range units, such as IFR2023A or similar, are barely sufficient.

Seriously consider hiring (briefly) a really high specification generator for your tests, such as an Agilent 8665, R&S SMHU or an IFR2040, for example.

The EN300-220 Class 1 radio specification is quite exacting, but there is no reason why a well-designed low-power module shouldn't comply with it. On the other hand, there are plenty of designs which are nowhere near this performance level in the marketplace and some of them are being sold as 'class1' units.

Good luck testing and, as always, "buyer beware"!

Myk Dormer is Senior RF Design Engineer at Radiometrix Ltd ***www.radiometrix.com***

NOTE: EN 300 220-1 V2.2.1 (2008-04) IMPLEMENTING 2009

http://webapp.etsi.org/workProgram/Report_WorkItem.asp?wki_id=27107
http://webapp.etsi.org/workProgram/Report_Schedule.asp?WKI_ID=27107

Lossless Feedback Amplifiers

Chris Trask goes into the theory, practice and advanced techniques of creating lossless feedback amplifiers. This article comprises two parts, published here in the next issue of *Electronics World* magazine.

AMONG THE TOPICS of particular interest to designers of telecommunications and radio equipment are intermodulation (IMD) and noise figures (NF) as they are limiting factors in the performance of radio receivers. The minimum detectable signal (MDS) of a receiver is determined by way of the NF and together with the third-order input intercept point (IIP3) determines the spurious-free dynamic range, commonly referred to simply as dynamic range.

The earliest stages of a receiver determine the overall dynamic range, and it is important that close attention be given to the design of the low-noise amplifier (LNA) stage(s) immediately after the antenna input terminals. The NF is very much a matter of proper device selection, together with proper device impedance matching and sufficient decoupling of power supply noise.

The IMD performance of an amplifier is very dependent upon the biasing conditions, as well as the linearity of the device itself. Unfortunately, the bias conditions that provide high IMD performance very often conflict with otherwise excellent NF performance, therefore, a judicious compromise needs to be considered in the architecture of the receiver system design so that the LNA design remains practical yet near optimal in terms of overall gain, NF, IMD and bandwidth performance objectives.

Series/Shunt Feedback Amplifiers

One method to obtain good IMD performance is to make use of negative feedback in the amplifier design. For RF amplifiers, two methods of negative feedback that find wide application are series/shunt feedback and lossless feedback.

The series/shunt feedback amplifier topology, shown in **Figure 1**, was devised and patented by Leonard Seader and James Sterett of Avantek. Shown in basic form in Figure 1, the amplifier consists of a transistor and two resistors. The design equations for the two resistors are quite simple, and the amplifier input and output impedances (R_{IN} and R_{OUT}, respectively) are:

$$R_{IN} = R_{OUT} = \sqrt{R_{FB} \times R_E} \quad (1)$$

The voltage and current gain are determined by way of:

$$\frac{v_{out}}{v_{in}} = \frac{i_{out}}{i_{in}} = \frac{1 - \sqrt{R_{FB}/R_E}}{2} \quad (2)$$

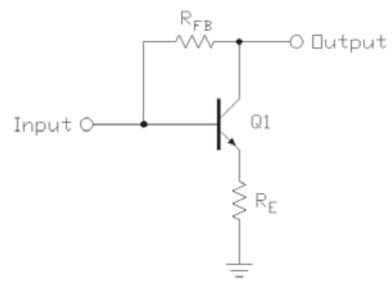


Figure 1: *Series/Shunt Feedback Amplifier*

and the amplifier voltage gain A_V is defined as:

$$A_V = \left| \frac{v_{out}}{v_{in}} \right| \quad (3)$$

from which the amplifier power gain is determined by:

$$G = 20 \log A_V \quad (4)$$

and which is also used to derive the values of the two resistors

$$R_{FB} = R_L \left(2 A_V + 1\right) \quad (5)$$

by way of:

$$R_E = \frac{R_L^2}{R_{FB}} \quad (6)$$

N	M	Rc Ohms	Gain dB
1	2	150	6.02
5	3	400	9.54
11	4	750	12.04
19	5	1200	13.98

Table 1: *Lossless Feedback Amplifier Transformer Ratios*

– Part 1

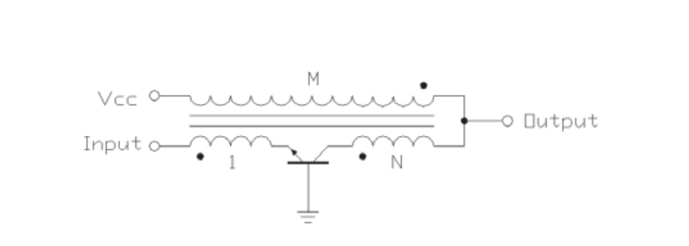


Figure 2: *Lossless Feedback Amplifier*

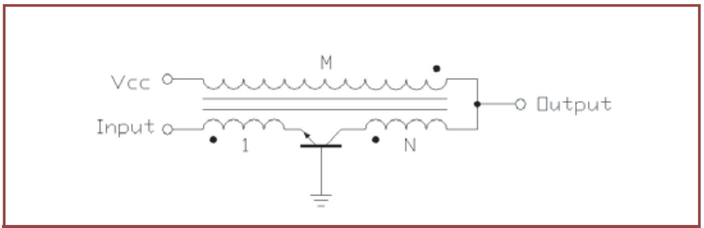


Figure 3: *Basic Linearity Augmentation*

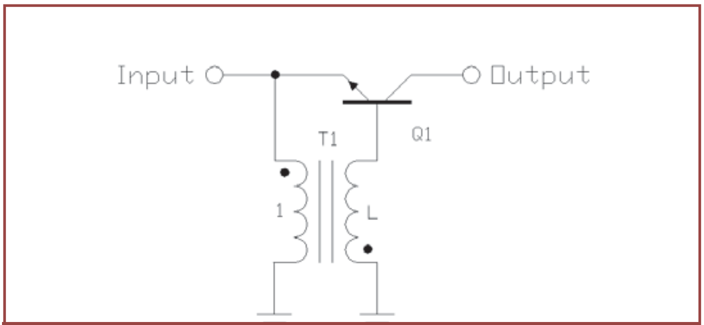


Figure 4: *Passive Augmentation*

Although the series/shunt amplifier may be convenient, the use of resistors as feedback elements, whether carbon or metal film, is detrimental to otherwise good NF performance, and the overall NF of the series/shunt amplifier is often many dB beyond the NF of the transistor.

The series/shunt amplifier can provide a good level of saturated power since the collector is coupled directly to the load. However, the degree of linearization is dependent upon the difference between the signal gain of the transistor and the closed loop gain. As a general rule of thumb, the transistor signal gain should be 3dB or greater than the closed loop amplifier gain.

An interesting innovation in this topology is to use a cascode pair in place of the single transistor. Although this configuration decreases the amplifier saturable power, it does increase the high cutoff frequency.

Lossless Feedback Amplifiers

Perhaps the single most significant development in high dynamic-range amplifiers has been that of the lossless feedback amplifier. Conceived and patented by David Norton and Allen Podell of Adams-Russell, this topology is often referred to as a Norton amplifier and, occasionally, as noiseless feedback. These amplifiers have seen a wide usage amongst radio amateurs, telecommunications and radio astronomers.

As shown in **Figure 2**, the lossless feedback amplifier consists basically of a transistor and a three-winding transformer. Overall, the performance is less dependent on the transistor and more dependent on the transformer, particularly the coupling coefficient between the three windings. Neglecting the finite emitter input resistance and the induced losses and less than unity coupling of the transformer, the amplifier gain is simply:

$$G = 20\,log\,M \tag{7}$$

and the input impedance of the amplifier is:

$$R_{IN} = R_L \frac{M + N + 1}{M^2} \tag{8}$$

where R_L is the load resistance. For an amplifier whose input impedance is equal to the load impedance, the turns ratios of M and N are related by:

$$N = M^2 - M - 1 \tag{9}$$

An important factor to take into consideration is the collector load resistance, which is:

$$R_C = R_L(M + N) \tag{10}$$

The collector load resistance R_C limits both the saturable power and the high frequency cutoff of the amplifier, and is rarely given much attention. In **Table 1** it can be readily seen that as the turns ratios of the transformer are increased to obtained higher gain, R_C rises rapidly, depriving the amplifier of both dynamic range and bandwidth, as we shall see shortly.
With respect to the saturable power, the maximum peak power that can be obtained is a result of R_C and the combination of the quiescent collector-emitter voltage (V_{CE}) and the collector saturation voltage ($V_{CE\ SAT}$) of the transistor: the collector saturation voltage ($V_{CE\ SAT}$) of the transistor:

$$P_{max} = \frac{(V_{CE} - V_{CE(SAT)})^2}{R_C} \tag{11}$$

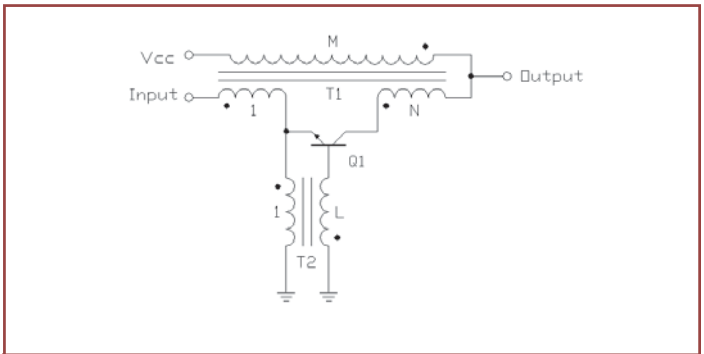


Figure 5: *Lossless Feedback Amplifier with Passive Augmentation*

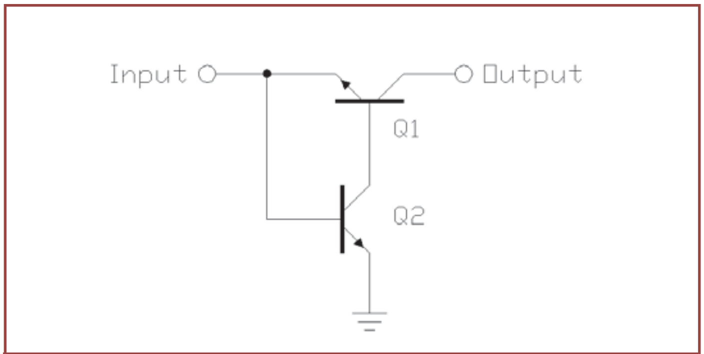


Figure 7: *Active Augmentation*

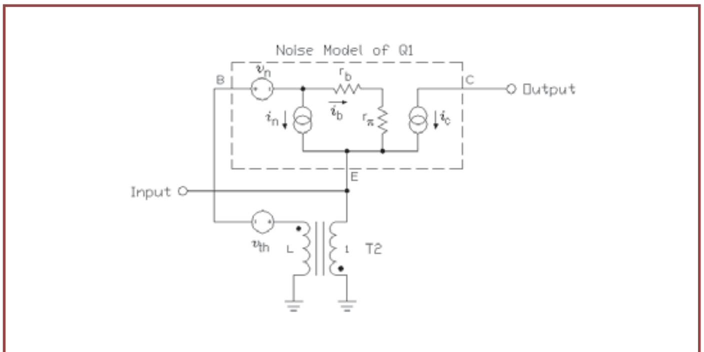


Figure 6: *Noise Model of Passive Augmentation*

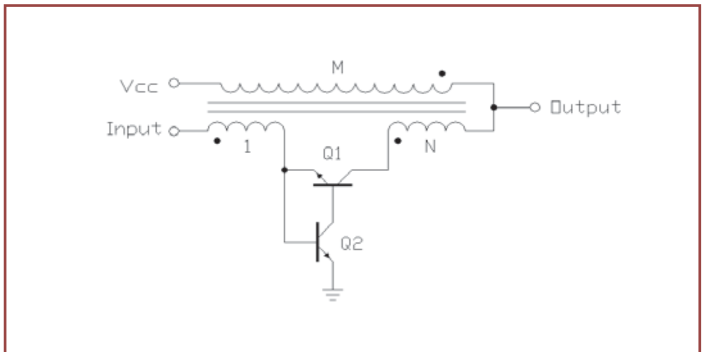


Figure 8: *Lossless Feedback Amplifier with Active Augmentation*

and the average power, which is what we typically measure, is 3dB less than this.

The high cutoff frequency dependency on R_C is a result of the collector-emitter and collector-base capacitances of the transistor (C_{oss} and C_{rss}, respectively) as well as the net capacitance across the output windings of the transformer (C_{NM}):

$$f_{max} = \frac{1}{R_C(C_{oss} + C_{rss} + C_{NM})} \qquad \textbf{(12)}$$

Other factors, such as the leakage inductances of the transformer and stray capacitance to ground, serve to further lower f_{MAX}. The transistor capacitances of **Equation 12** in conjunction with R_C represent a serious obstacle to higher frequency designs, and variations of the basic lossless feedback topology, to be discussed later, may need to be brought to bear.

These last two equations speak volumes concerning the selection of the amplifier transistor. Although common-base amplifiers are generally considered to be more linear than their common-emitter counterparts, they are still subject to the consequences of poor saturation characteristics. If the designer is concerned about overall power efficiency in conjunction with good linearity, then the $V_{CE\ SAT}$ characteristics of the transistor become an important issue in device selection. Not only is the saturation voltage itself to be considered, but the transfer characteristics should be such that the approach to saturation is a sharp knee rather than a smoothed curve.

Transistors that match this description include the 2N2222, 2N2907, MPS6512, MPS6513 and the 2N3553, all of which are suitable for frequencies from VLF to low VHF. For higher frequencies, devices having lower capacitances such as the NE681 series are worthy of consideration.

An additional factor that limits the overall performance of the lossless feedback amplifier has to do with the approximation that Norton and Podell made in formulating the design equations, which was the assumption that the emitter input resistance (R_E) of the transistor is negligible. This is, of course, very convenient in both the overall theory and the design process, but in reality R_E is not only finite but it is also nonlinear and both of these qualities detract from the overall potential linearity of the amplifier.

We could, or course, increase the transistor bias current so as to reduce R_E, but this would have other impacts on the amplifier. First, it would increase the supply current and, therefore, reduce the amplifier power efficiency. In addition, transistors typically have their best NF and linearity characteristics at lower collector currents.

Linearity Augmentation

Since the issuance of the initial patent, little has been done to continue the further development of the lossless feedback amplifier, which is a bit surprising given the emphasis on raising the signal-to-noise ratio (SNR) and IMD performance of telecommunications equipment so as to adequately get above the noise floor and further improve the overall dynamic range. To this effect, linearity augmentation was devised and patented as a means of overcoming the intrinsic finite and nonlinear R_E that compromises the performance of common-base and lossless feedback amplifiers, as well as to improve the NF. Basically, as shown in **Figure 3**, the signal voltage at the emitter of the transistor is sensed by an inverting amplifier and the output is applied to the base. For a simple voltage amplifier having a voltage gain of A_V, the reduction of R_E is:

$$R'_E = \frac{R_E}{A_V + 1} \qquad \textbf{(13)}$$

where the emitter resistance R_E is determined by the familiar relationship:

$$R_E = \frac{V_{BE}}{I_0 \, å^{\frac{q V_{BE}}{k T}}} \tag{14}$$

Since wideband voltage amplifiers having simultaneously high gain and low noise at RF frequencies are somewhat impractical to achieve, two alternative methods were devised, being passive and active augmentation. With passive augmentation, shown in **Figure 4**, the voltage amplifier is replaced with a simple two-winding transformer having an arbitrary windings ratio of L. With this method, the reduction of R_E becomes approximately:

$$R'_E = \frac{R_E}{\left(1 - \frac{L}{h_{fe}}\right) å^{(L+1)}} \tag{15}$$

For transistors having h_{fe} of 100, a reduction of R_E in the order of 95% is realizable with a transformer having a turns ratio of 1:3. **Figure 5** illustrates a lossless feedback amplifier with passive augmentation.

Passive augmentation also has benefits in terms of NF. Using **Figure 6** as a reference, the noise of the amplifier transistor Q1 passes through the primary winding of the transformer T2, where the voltage is amplified and inverted across the secondary winding and then coupled to the base of Q1, resulting in a reduction of noise voltage by approximately 1/L.

The noise source v_{th} represents the noise added by the bulk and induced losses as well as the Barkausen noise of the core of T2. Due to the nature of transformers, this reduction is limited to the thermal, or Nyquist noise of the transistor as the 1/f, or "flicker" noise will most likely be below the low cutoff frequency of the transformer.

There are, of course, practical limitations to the windings ratio L of the augmentation transformer. As L is increased, parasitics such as leakage inductances and both intra- and interwinding capacitances can limit the frequency bandwidth. To further improve the reduction of R_E, active augmentation (as shown in **Figure 7**) can be employed. Here, the second, or augmentation transistor Q2 amplifies and inverts the signal voltage at the emitter of the amplifier transistor Q1, resulting in a substantial reduction in R_E:

$$R'_E = \frac{R_{E2}}{\left(h_{fe1} + 1 + \frac{1}{h_{fe2}}\right)} \tag{16}$$

where R_{E2} is the emitter resistance of transistor Q2 and h_{fe1} and h_{fe2} are the signal current gains of transistors Q1 and Q2, respectively. A schematic depicting the application of augmentation in a lossless feedback amplifier is shown in **Figure 8**.

Provided that Q2 is properly biased and matched for its best noise performance, the use of active augmentation provides a substantial reduction of the 1/f noise of the amplifier transistor Q1, but at the same time will add some amount of thermal noise.

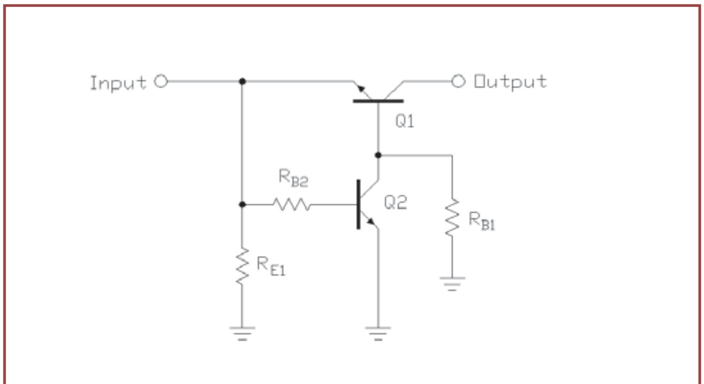


Figure 9: *Noise Matching for Active Augmentation Transistors*

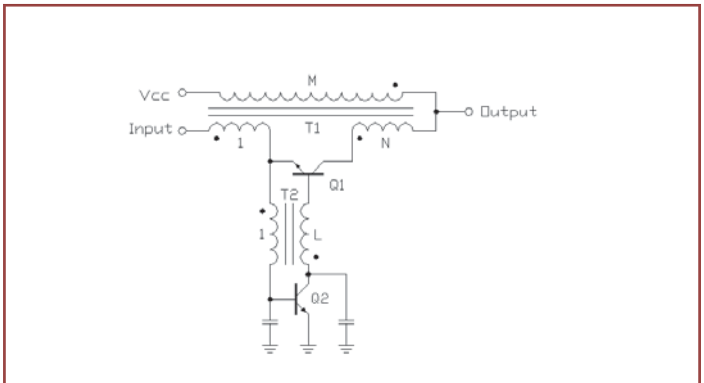


Figure 11: *Lossless Feedback Amplifier with Tandem Augmentation*

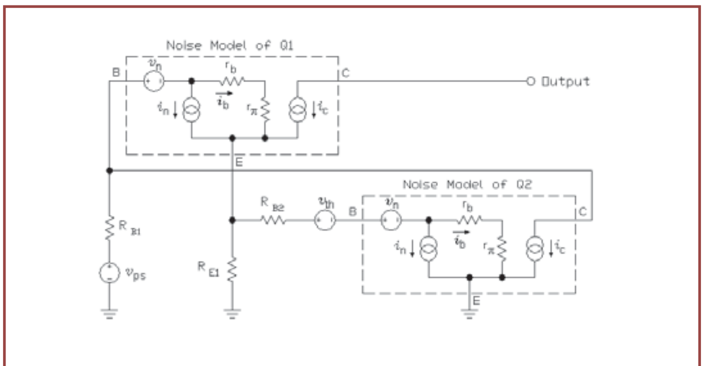


Figure 10: *Detailed Noise Model of Active Augmentation*

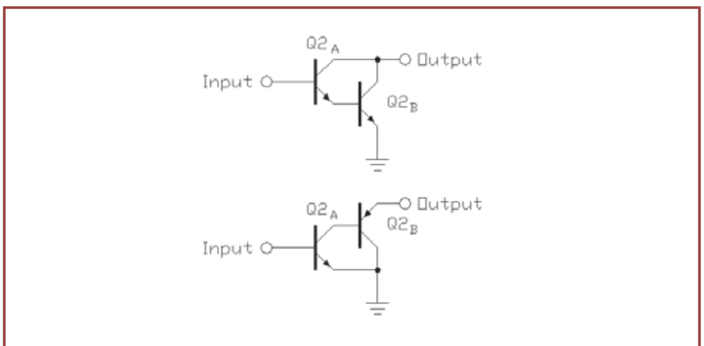


Figure 12: *Darlington (upper) and Sziklai (lower) Compound Transistor Pairs*

Referring to the low frequency schematic in **Figure 9** and the detailed noise model in **Figure 10**, the resistor R_{B1} is the biasing resistor (or resistors) shared by the collector of Q2 and the base of Q1. Resistor R_{E1} is the emitter biasing resistor of Q1, and resistor R_{B2} is the base resistor added between R_{E1} and the base of Q2 to properly match Q2 for its best noise performance.

In the detailed noise model of Figure 10, the noise source v_{ps} represents the noise from the power supply, and the noise source v_{th} represents the thermal noise added by the three resistors. Similar methods of noise reduction have been described before for common-emitter amplifiers and microwave oscillators.

An added benefit of active augmentation is the reduction of even-ordered IMD products, such as 1x1, 2x2, 3x3, etc that exist at low or baseband frequencies and which are rarely, if ever, mentioned in the overall scheme of improving IMD performance. The transformer used with passive augmentation might not conduct these signals as they would very likely fall below the low cutoff frequency of the transformer. However, in an amplifier using direct-connected active augmentation these signals would be corrected down to DC.

Both passive and active type augmentations have benefits and shortcomings. Passive augmentation provides the needed reduction in the emitter resistance and some reduction in the thermal noise, but due to the nature of transformers it has little effect on the even-order baseband IMD products or the low frequency 1/f noise. Active augmentation provides a higher degree of emitter resistance reduction plus effective reduction in 1/f noise and baseband IMD products, but at the same time has the potential to add thermal noise.

The two methods of augmentation described thus far may be combined so that their benefits may complement each other, thereby overcoming each others' shortcomings. This method is known as tandem augmentation, and is illustrated in the schematic of **Figure 11**. Here, the transformer T2 provides the inband passive augmentation to reduce inband IMD products along with some reduction in the thermal components contributing to the NF, while the transistor Q2 provides the baseband augmentation to correct the baseband IMD products as well as the low frequency 1/f components of the NF. The bypass capacitors at the base and collector of Q2, together with transformer T2 create a crossover network that determines the point at which the passive and active augmentation dominate.

The application of tandem augmentation presents an interesting opportunity to dramatically lower the NF of the amplifier. Returning for the moment to the detailed noise model of Figure 10, the overall benefits of lowering the 1/f noise of the amplifier by way of the augmentation transistor Q2 may be exploited by making use of audio frequency transistors that have low NF together with high gain, and transistors having these characteristics include such devices as the 2n2484, 2N5089, MPSA18 and the BC549C. With collector currents of the order of 1.0mA, these devices have NF in the order of 2dB, and monolithic Darlington pair transistors such as the MPSA14 and the BC517 have similar NF along with higher gains in the order of 10,000 or more.

The NF of the previously mentioned low-noise audio transistors may be reduced to as little as 0.5dB or less when the collector current approaches 10μA, however this may not be sufficient for providing proper biasing of the amplifier transistor Q1. In such circumstances, the designer may choose to combine these devices with a second transistor to form a discrete Darlington or Sziklai pair, as shown together in **Figure 12**.

Here, the first transistor Q2A is operated at the very low collector-current commensurate with the desired low NF, while the second transistor Q2B is operated at a higher collector current that is appropriate for maintaining the proper biasing condition for the amplifier transistor Q1. As an example, a BC549C can be combined with an MPS2222 or MPS2907 to make either a Darlington or Sziklai pair, respectively, that will have a NF of approximately 0.55dB and a current gain in excess of 20,000, which surpasses the performance available from monolithic Darlington pairs. Making use of such a combination for the augmentation transistor can result in amplifiers whose NF is below 1dB.

Very often, the amplifier transistor Q1 requires a fairly low source-resistance in order to be properly matched for low NF. In such cases, a single augmentation transistor or a Darlington pair in combination with the biasing resistor R_{B1} of Figure 9 may not be able to provide the needed low source resistance. However, the Sziklai pair, shown in Figure 12, has an emitter follower output and can easily provide the needed low resistance, and the designer only needs to place an additional resistor in the circuit of Figure 9 to achieve the source resistance required by the amplifier transistor Q1.

The second part of this article will be published in the next issue (April cover month) of Electronics World magazine. Order your digital copy of that issue on-line at **www.electronicsworld.co.uk**

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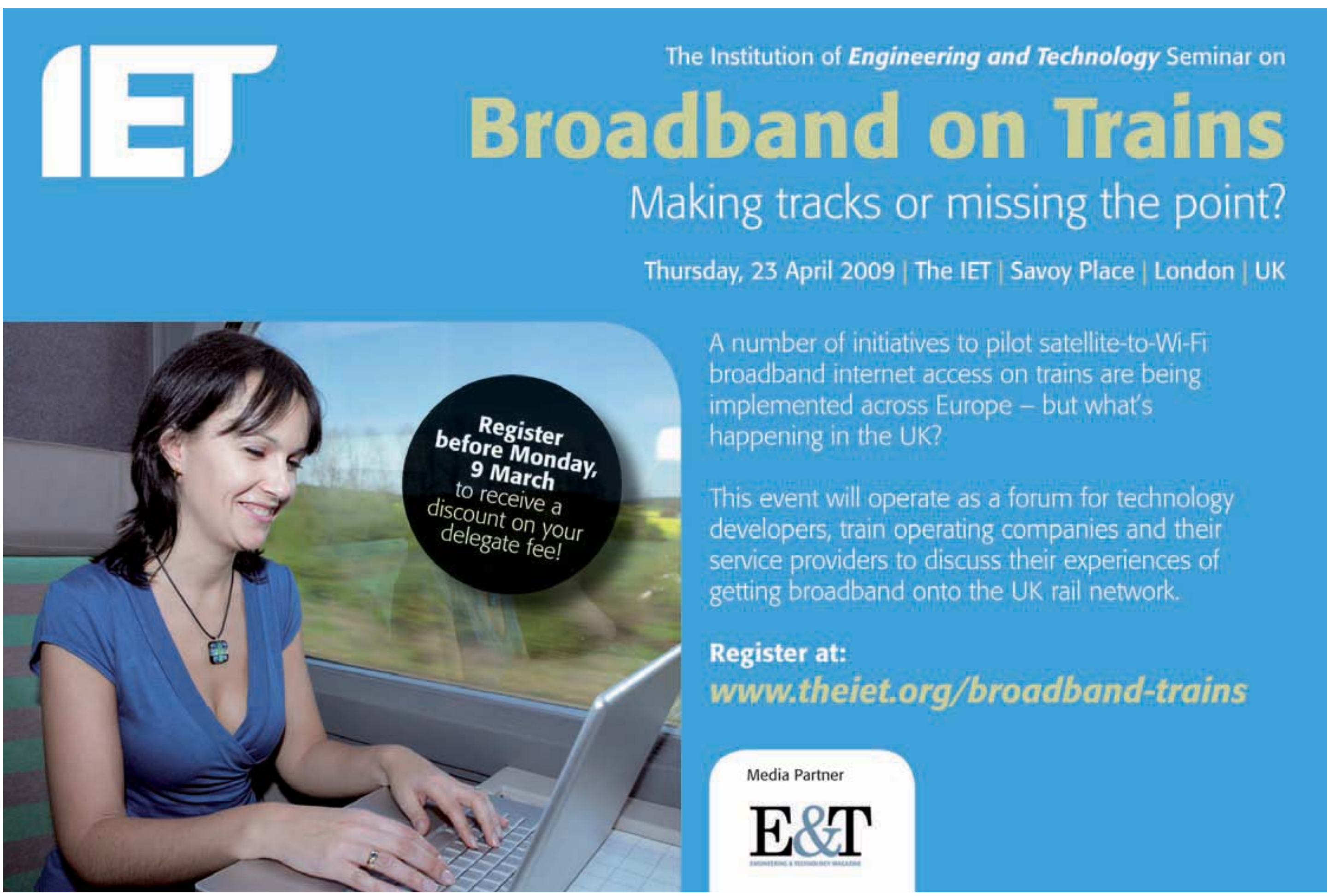

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The Hum Figure

THE FRUSTRATION that immediately follows noise and hum (H&N) that may come out of the loudspeaker-amplifier chain in the gap between music titles begs the question of "how to measure or calculate the hum artefact content of the output noise of a power amplifier with input shorted and output loaded with the rated load".

Gaps between CD music titles or from other low-noise sources should be hum and noise-free. There could be one exception: vinyl records plus phono-amp. In most cases their H&N level "marches" 1:1 through the whole amp chain, thus, producing the Signal-to-Noise (SN) ratio at the phono-amp's output without major increase at the output of the PA as well.

Those customers who are not familiar with the in-depth electronics of a power amplifier (PA) or an integrated amplifier (pre plus PA) but want to buy one, are confronted and swamped by a lot of different information, which at some degree can even be contradictory. The information does not always help to answer the rather simple question as above.

Equally, I can imagine that not many salespeople will have a real idea of how to answer this question. I bet something like "concerning H&N the A-weighting measurement method of the output noise sufficiently indicates what's going on" could be a practical excuse. A test in a sales shop becomes an interesting adventure!

Unfortunately, up to today, nobody could kill this (and in my eyes silly) A-weighted noise measurements for amplifiers and their representation in advertising, marketing papers and test magazines. To bring more into line with the equivalent loudness contours of the human ear, A-weighting (NAB/ANSI S1.4-1986) became a standard for use in sound measurements only. **Figure 1** shows the A-filter measurement effect on a hum-infected valve PA. Whereas a purely white noise signal is dampened by only 2.05dB in B_{20k} = 20Hz …20kHz (**Figure 2**). A hum infected output signal of a PA produces significantly higher differences of non-weighted SN_{ne} versus an A-weighted output $SN_{ne.a}$.

Nevertheless, in test magazines and manufacturer prospectus papers in most cases noise of amps is indicated in dBA with the following wording: Noise or Signal-to-Noise-Ratio (SN or SNR) = 79dB (A) or 79dBA. One reason might be the fact that the measured A-value looks better (bigger) than that of a non-weighted SN measurement. In general, this is not a reason against the use of A-weighting but a reflection of the brochure's lack of depth.

Thorough Information is Necessary

Any SN significantly needs additional information for proper use: the noise measuring bandwidth as well as a reference level to hang on the SN (which in fact should have a negative meaning because, normally, in the audio environment noise, voltages are smaller than the chosen reference level).

Furthermore, the information of the input and output loads of the amp under test are equally necessary to know. Otherwise, a simple "79dBA" does not tell anything – unless they want to cover something up.

In addition, and because of the following arguments, A-weighting measurements are not favourable and they are not telling the truth on amp output H&N at all. If all of this information was given, the value of say "-79dBV(A)" may hide from extremely different noise backgrounds which can only be detected by non-weighted SN measurements, thus, leaving the customer very unclear about the amp's real H&N situation.

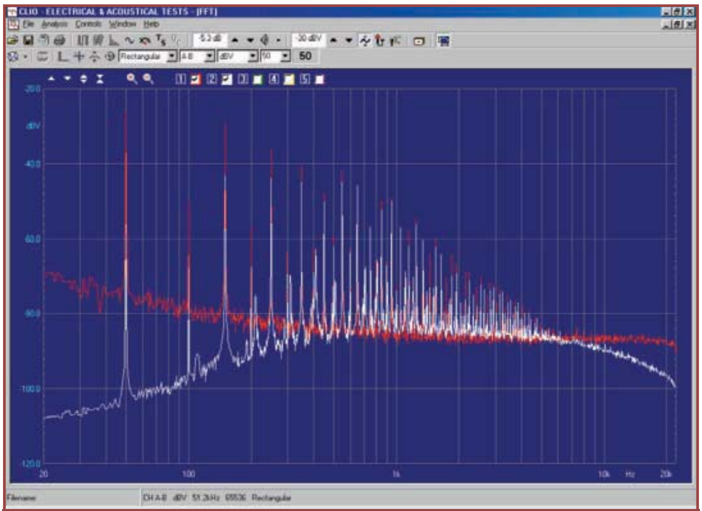


Figure 1: *Valve PA (R channel) H&N output non-weighted (red) and A-weighted (white)*

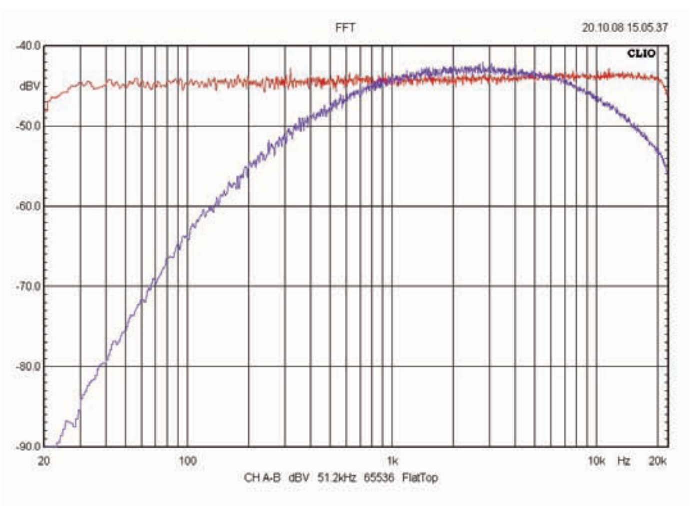


Figure 2: *0dBV white noise (red) and weighted by an A-filter (blue)*

The first case might be an amp without any hum artefacts in its output noise voltage; therefore, its unweighted output Signal-to-Noise-Ratio SN_{ne} would become (-79dBV(A) + 2.05dB)) = -76.95dBV. The second case might be an amp with heavy hum artefacts in its output noise voltage (see Figure 1). Therefore, the amp's non-weighted SN_{ne} becomes (measured) or -67.8dBV = (-79dBV(A) + 2.05dB + 9.15dB). Here, the hum counts at least for 9.15dB in the chosen A-weighting environment, more or less a value quite often found in the output H&N voltage of valve power amps or lousy developed solid-state amps. However, in reality, the hum content of the output H&N signal is far bigger, app. 21dB (**Table 1**, F-10).

Burkhard Vogel discusses the creation of a specific figure as a measure of mains induced hum at the output of power amplifiers

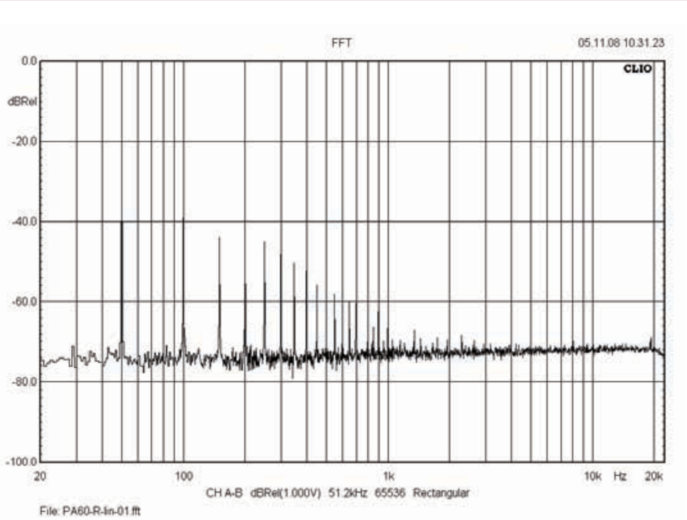


***Figure 3:** PA1 output H&N (BJT 30W/4R, $0dB_{rel}$ = -66dBV, R channel)*

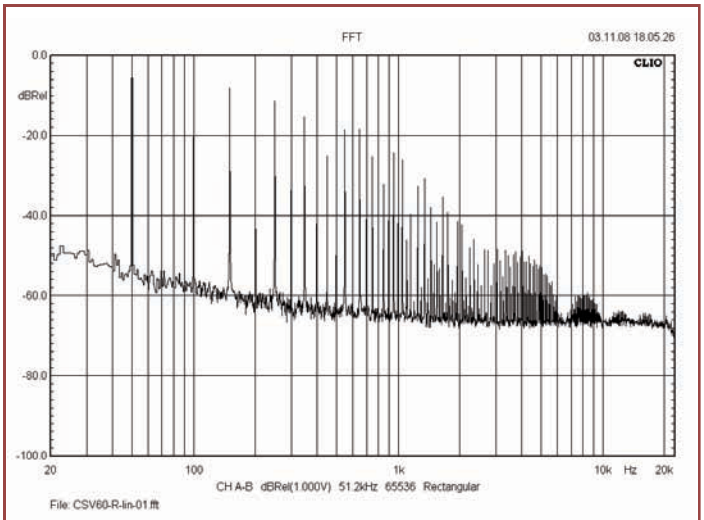


***Figure 4:** PA2 output H&N (valve 30W/4R, $0dB_{rel}$ = -66dBV, R channel)*

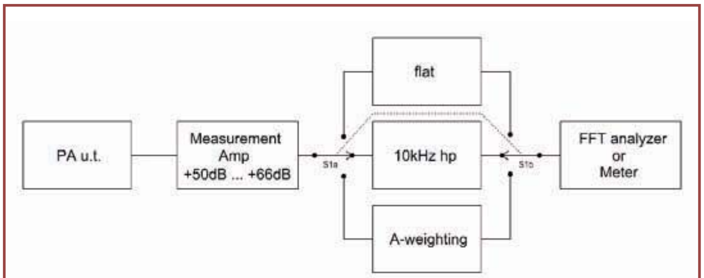


***Figure 5:** General HF_e measurement setup*

Standardizing Information

So, what does 79dBA say? I guess nothing for further H&N examination and nothing about the amp's real H&N quality.

Concerning output noise of PAs more truth would be in the table if all manufacturers could be convinced to indicate a standardized set of information, such as:

1. Treatment of the input: shorted or otherwise loaded;
2. Treatment of the output: loaded with the rated load impedance "I", i.e. I = 4R or 8R or any other I;
3. Selection of bandwidth B: for example, B = B_{20k} = 20Hz… 20kHz or any other B
4. Selection of the reference level: for example 1V [0dBV] at rated output load or any other reference level like the one of the studios: 0 or +6dBu;
5. SN_{ne} measurement result expressed in [-x dBV] provided that 4 have been taken into consideration ($_{ne}$ = non-equalized = flat frequency response in B_{20k});
6. HF_e measurement or calculation result expressed in [y dB] (X_e = 20log(X)).

"HF_e" is a new term. I call it the Hum Figure and its unit is dB. It is the result of further developments with Douglas Self's 400Hz hp filter method to measure SNs without the low-frequency influence of 1/f noise and hum, described in his *"Self on Audio"* book (Precision Preamplifier '96 Part I, ISBN 0 7506 4765 5)). HF_e indicates the amount of hum (mains fundamental plus its entire harmonics) as a part of the whole output H&N voltage of a PA. The calculation is as follows ($SN_{ne.wn}$ is the SN based on the flat white noise ('wn') level of the amp):

$$HF_e = SN_{ne} - SN_{ne.wn} \qquad (1)$$

In nearly all cases, PAs produce a rather flat white noise (wn) in B_{20k}. With corner frequencies up to 1kHz exceptions come from the 1/f noise at the lower end of B_{20k} (see **Figure 4**). Because of its rather small impact on HF_e, I will ignore it in the following paragraphs (more on the calculation of 1/f noise: see the author's **"The Sound of Silence",** Springer 2008, ISBN 978-3-540-768 83-8).

Demonstrating the Usefulness of HF_e

To demonstrate the usefulness of the HF_e method I have chosen two very different example PAs. PA1 is a self-developed solid state BJT driven 2 x 30W/4R PA, extremely low-noise and with a very low hum contamination.

PA2 is a valve-driven 2 x 30W/4R BRAUN CSV60 (ECF80+2xPL509), a high-end product from the 60's, extremely hum contaminated, thus, not very low in noise at all. Figures 3 and 4 allow studying their output H&N voltages. The wn at approximately 10kHz differs by approximately 6dB only!

To get $SN_{ne.wn}$ I will propose two different approaches. The first one (= filter approach = F-A) uses a special hp filter method, whereas the second one (= FFT approach = FFT-A) works with FFT analysis and some additional calculations. **Figure 5** shows the measurement setup. The results of both approaches got summed-up in Table 1. Result differences became rather small: within 1dB only.

The idea behind these approaches is that, if there would be no hum artefacts, the value of the amp's SN_{ne} would become automatically $SN_{ne.wn}$ (any tiny difference will indicate a certain 1/f noise effect). Therefore, besides a more truth generating SN_{ne}, in addition, HF_e would become a very strong PA design quality signal for the customer.

First HF_e evaluation approach: F-A

Assumed that there is no FFT analyser at hand, a simple dB calibrated millivolt meter with a 20kHz lp at its input will work as well, if the following 10kHz hp filter plus +60dB low-noise measurement amplifier in front of it is used. Described in detail in my *"The Sound of Silence"* book, I performed all measurements with my own measurement setup. **Figure 6** shows the circuitry of the Chebyshev 0.1dB 10kHz hp. The measurement amp is created with an extremely low-noise input op-amp like the LT1028 or the AD797.

To get good measurement results and to overcome the noise of the filter at least +50dB gain of the measurement amp must be placed in front of the hp. I have chosen a gain of +54dB in front and +12dB after the hp, hence, in total +66dB. To get the final SN values subtraction of the total gain of the measurement amp from the measured value at the meter is essential.

The hp filter block consists of an output amp (OP8) that allows to trim the gain (P8) of the whole block to a specific value (e_{out}/e_{in} = approx +3dB). The theory behind is because the wn in a B_{20k} frequency band was reduced by approx 3dB when filtered with a 10kHz hp, plus 20kHz lp to create a frequency band of 10kHz = B_{10k}. The reduction will be exactly 3dB with a filter slope of >> 100dB/octave but, together with the anti-aliasing filter of the FFT analyser or the 26kHz sixth order Butterworth lp of my own measurement equipment (-0.33dB/20kHz), the chosen filter slopes work sufficiently well. I have chosen this frequency band because of its total cleanness of hum artefacts.

Fed with a broadband white noise signal and with S3 set to "Fi on", we have to trim the gain of the filter block (P8). OP8's output voltage must have the same value like the one in "Fi off" mode. Thus, with filter "on" we get $SN_{ne.wn.pa}$, the wn based SN of the PA. Exact trimming is essential, otherwise, a PA without or with only tiny hum artefacts might create a negative HF_e.

Figure 7 shows the result of the trimming efforts. The delta between the noise level at 0Hz-2.5kHz and 10kHz-20kHz represents the measurement amp's gain in front of the hp.

Second HF_e evaluation approach: FFT-A

For the FFT analysis in B_{20k}, the division of the FFT size by the sample rate should be ≤ 1Hz. Because of Nyquist, the sample rate must be > 2*20kHz. With the measured SN_{ne} and picked out FFT values the calculation of $SN_{ne.wn}$ goes like this: the -dBV value of the white noise voltage $e_{wn.10k.e}$ can be picked out in the hum-free white noise region between 10kHz and 20kHz. To get $SN_{ne.nw}$ we simply have to add the dB-equivalent $B_{20k.e}$ for the bandwidth B_{20k} (see **Figure 8** and **Equation 2**).

$$B_{20k.e} = 20 * \log(\sqrt{B_{20k}}) = 43.006\text{dB} = 43\text{dB rounded}$$
$$SN_{ne.wn} = e_{wn.10k.e} + 43\,\text{dB} \qquad (2)$$

Depending on any result of FFT size divided by a sample rate that creates a solution >1Hz by halving of the sample rate the Figure 8 wn floor will move up in 3dB steps, thus reducing $B_{20k.e}$ by the same amount. For deeper details on these mechanisms, see Audio Precision's *"Audio Measurement Handbook"*, written by Bob Metzler.

With the input shorted, output loaded with 4R, bandwidth = B_{20k} and any 1/f noise voltage effect ignored the measurement and calculation results rounded to one digit after the decimal point are given in Table 1, lines 6-10 and will look as follows:

1 / A	B	C	D	E	F	G	H
2		SN data acquisition method	type of measurement or calculation	PA1-R	PA2-R	line unit	Remarks
3	(1)	F-A: measured SNs	$SN_{ne.B20k.meas}$ = lin in B_{20k}	-94.7	-67.8	dBV	
4			$SN_{ne.wn.20k.meas}$ = 10kHz-hp "on"	-95.5	-90.9	dBV	
5			**$Hf_{e.pa.meas}$ = lines (3-4)**	**0.8**	**23.1**	dBV	
6	(2)	FFT-A: wn picked out from FFT to calculate $SN_{ne.wn.20k}$	$e_{wn.10k.e}$ (wn at 10kHz picked out)	-72.0	-68.0	dBV	FFT averaging only 50, thus, picked out values of line 6 were rounded to 1 digits after the decimal point
7			gain of measurement amp	-66.0	-66.0	dBV	
8			$B_{20k.e}$ (wn equivalent in B_{20k})	43.0	43.0	dB	
9			$SN_{ne.wn.20k.calc}$ = lines(6+7+8)	-95.0	-91.0	dBV	deltas between lines 4 and 9 within 1dB max.
10			**$Hf_{e.pa.calc}$ = lines (3-9)**	**0.3**	**23.2**	dBV	deltas between lines 5 and 10 within 1dB max.
11	(3)	A-weighting	$SN_{ne.a.pa}$	-98.2	-79.0	dBV	

Table 1: *Results of first and second HF_e evaluation approach*

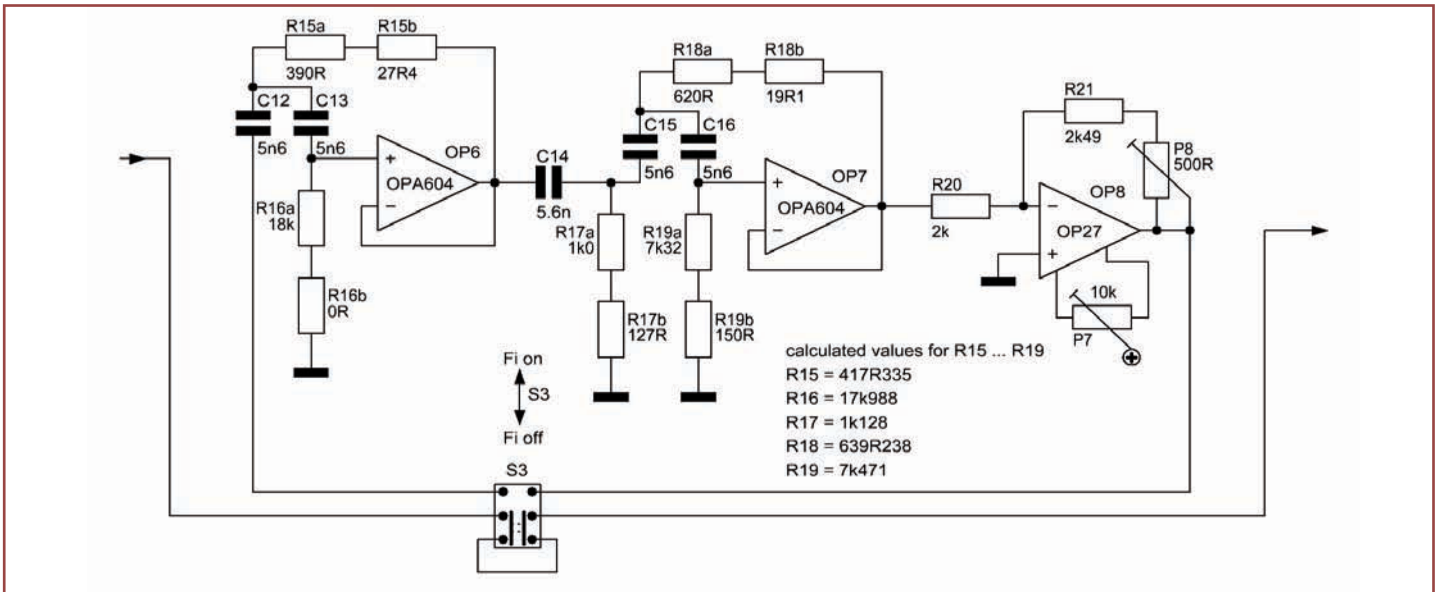


Figure 6: *Chebyshev 0.1dB, 10kHz, high-pass filter*

$$HF_{e.pa1} = SN_{ne.pa1} - (e_{wn.pa1.10k.e} + B_{20k.e})$$
$$= -94.7\,\text{dBV} - (-138.0\,\text{dBV} + 43.0\,\text{dB}) \quad (3)$$
$$= 0.3\,\text{dB}$$

$$HF_{e.pa2} = -67.8\,\text{dBV} - (-134.0\,\text{dBV} + 43.0\,\text{dB}) \quad (4)$$
$$= 23.2\,\text{dB}$$

In other words, without hum (and maybe a tiny portion 1/f noise) PA1's or PA2's SN_{ne} would become 0.3dB resp. 23.2dB better. As such, no A-weighting figure will produce these H&N quality-defining findings.

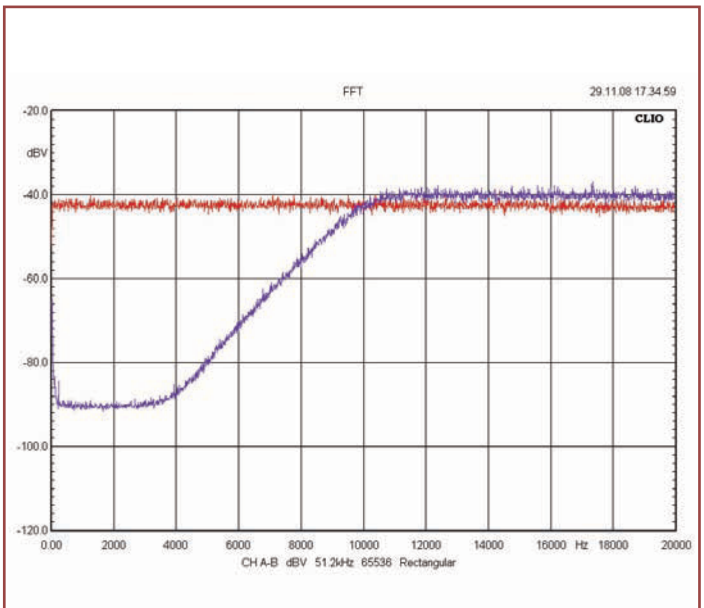


Figure 7: *10kHz hp weighting effect (blue) on 0dBV white noise (red and see Figure 8), thus generating 0dBV after careful trimming of P8*

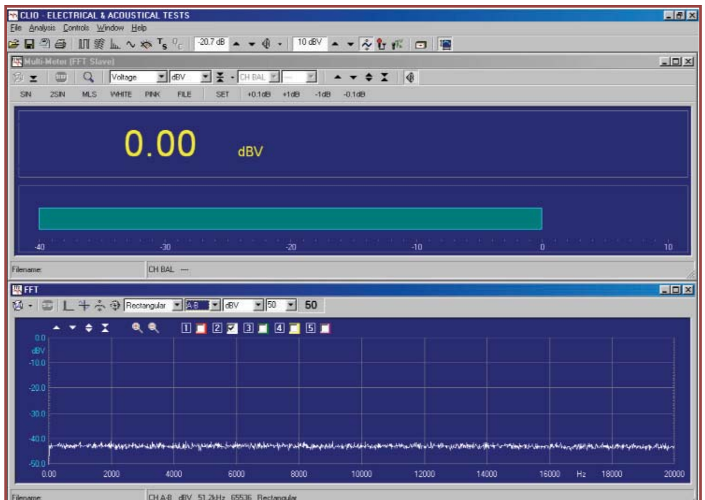


Figure 8: *$B_{20k.e}$ = 0dBV; white noise in B_{20k} with its voltage noise density line at -43dBV*

Summary

Like the PA's noise figure "NF_e" the hum figure HF_e does only give us the amount of hum that was added to a hum-free input signal. It does not give an absolute value of SN. This can only be done by measuring SN_{ne} or by rms add-up of all hum level values, plus the remaining wn.

However, both values, SN_{ne} and HF_e, give us enough hints to qualify a PA concerning the producer's efforts to create a low-noise and low-hum piece of value. Even today, this is essential to drive loudspeakers with high sensitivity, especially when listening to them at a rather short distance.

Depending on the volume potentiometer setting, any H&N contaminated input signal from a pre-amp or any other source adds to the PA's H&N signals. Thus, to get a full H&N picture of the whole amp chain connect the input of the PA to the output of the source and set the volume pot in "usual listening" or 11am positions.

Finally, and provided that H&N measurements were always be performed with the input shorted and within B_{20k}, the information about the H&N situation of a specific PA should be indicated by the manufacturers as follows:

$$SN / HF / I = -x\,\text{dB} + \text{reference level} / y\,\text{dB} / z\,R \quad (5)$$

Ear fatigue reduction in the

Wangeri Harrison Kaguongo analyses some of the approaches for reducing ear fatigue in audio amplifiers

AUDIO AMPLIFIER design is a multi faceted discipline that covers:

1. Power supply
2. Amplifier circuit:
   a. Input stage
   b. Voltage gain stage
   c. Current gain stage
   d. Protection
   e. Current sources/sinks/mirrors;
3. Speaker system; and
4. Audio Source.

Each has the potential to improve or ruin the listening experience. The power supply is the power base for the audio. If the power base/power supply is problematic or poorly designed, then it won't matter how good the amplifier design is.

Once the power supply is in place, the design of the speaker system can begin as well as the design of the audio source. In this article, we will cover the amplifier circuit

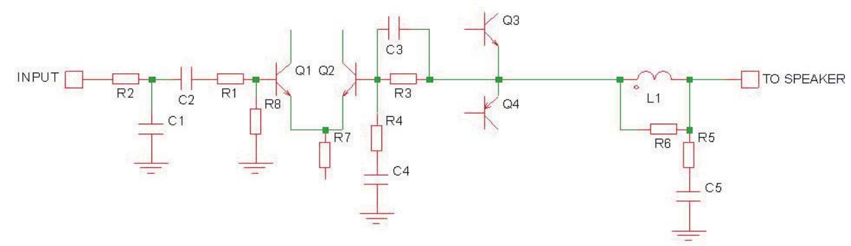


Figure 1: *One technique of eliminating ear fatigue and also radio frequency interference is to isolate C1 from the base of Q1 using a resistor R1*

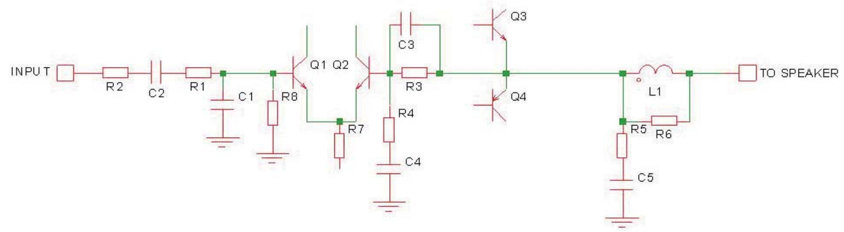


Figure 2: *Another technique is to have C1 at the base of Q1 and ensuring that C3 is present and that R5 and C5 are at the output of the power transistors just before R6 and L1*

Input

There are techniques of eliminating electromagnetic interferences as well as safeguarding against capacitor ear fatigue as shown in **Figures 1** and **2**. One way is to keep the input capacitor C2 as small as possible and to isolate it from the input transistor Q1 using a resistor R1. This goes hand in hand with keeping the capacitor C4 that shorts AC signals to ground on the feedback side of the LTP to the groundside, while the resistor R4 goes to the base of the input transistor Q2.

The small PF capacitors used for high frequency filtering C1, and also for high frequency feedback C3, can be directly attached to the transistor bases. For good interference rejection Figure 1 and Figure 2 show the relationship between the location of C1 in relation to the location of R5 and C5. But circuit topology and choices may

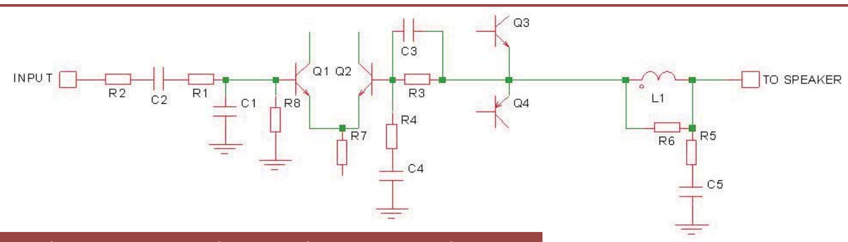


Figure 3a: *There are situations however that necessitate the placement of R5 and C5 after R6 and L1 for the sake of better amplifier stability, but this depends on the overall amplifier topology*

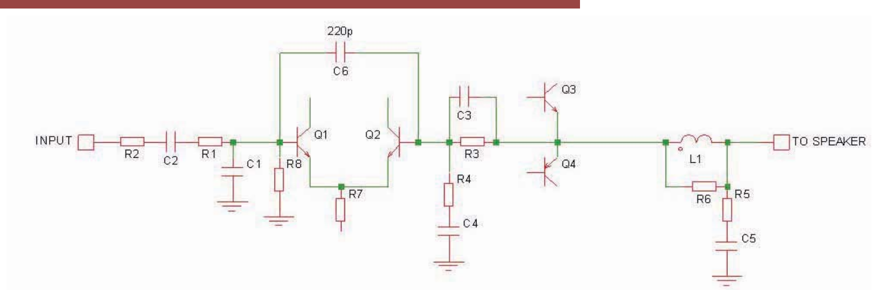


Figure 3b: *A 220pF capacitor across the bases of the LTP does no good and may destabilize the circuit that necessitates the placement of R5 and C5 after R6 and L1 for the sake of better amplifier stability, but this depends on the overall amplifier topology*

design of AUDIO AMPLIFIERS

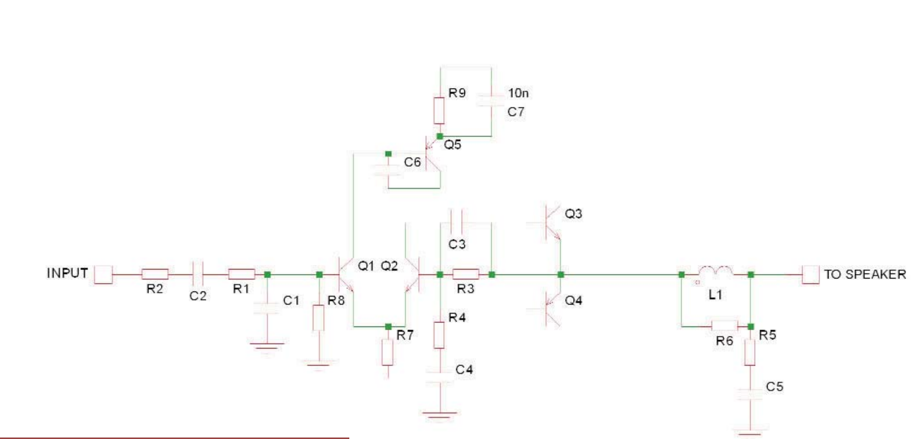


Figure 4: *10nf capacitor C7 is necessary to increase amp stability so long as C6 or C3 are present*

necessitate one to put R5 and C5 after L1 and R6. This will yield better circuit stability as a result of choices made in the topology of the input and output as shown in **Figure 3a**.

One interesting thing about amplifiers is that you may find the signal at the collector of Q2 distorted at certain frequencies, depending on the overall circuit topology. In relation to Figure 1, the capacitors at the input C2 and feedback C4 do affect the transient phase at low frequencies. The maximum recommended value for C2 is around 10uF, while that for C4 is around 100uF. The issue about capacitors in amplifier

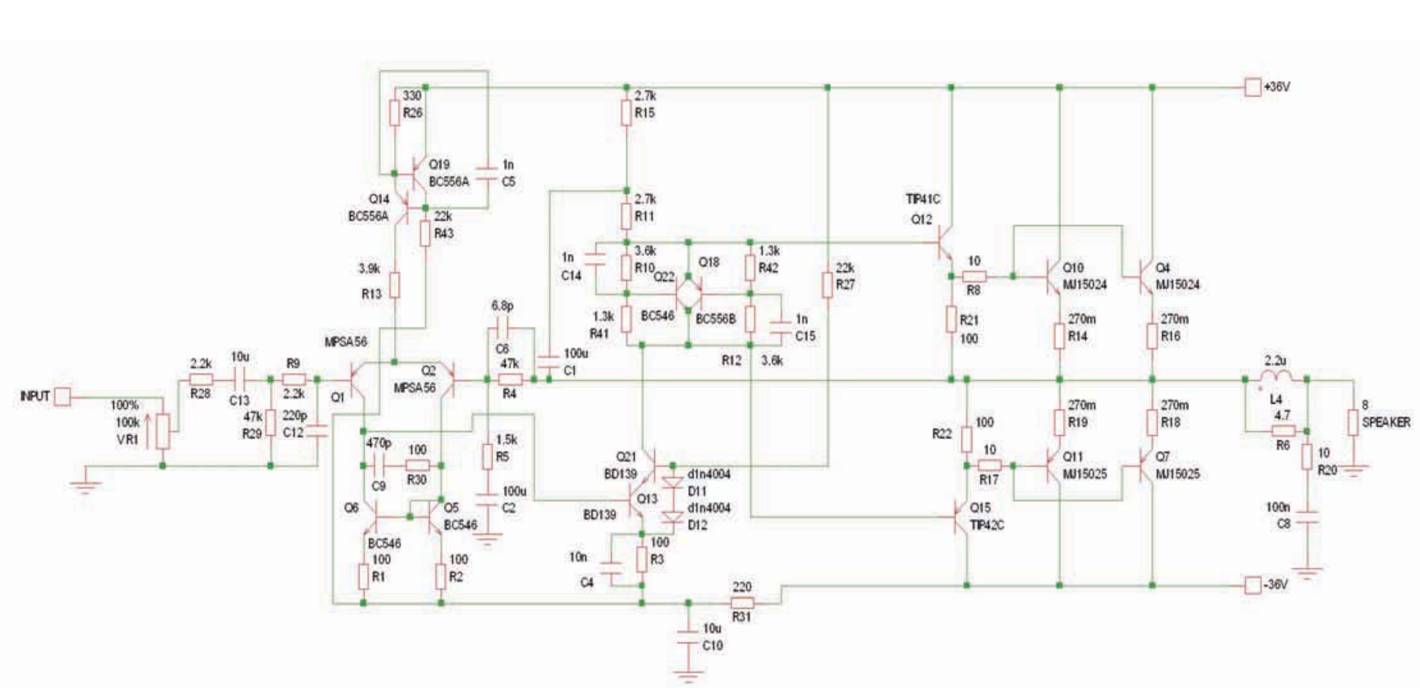


Figure 5: *A fully simulated and tested circuit that may be fun for designers*

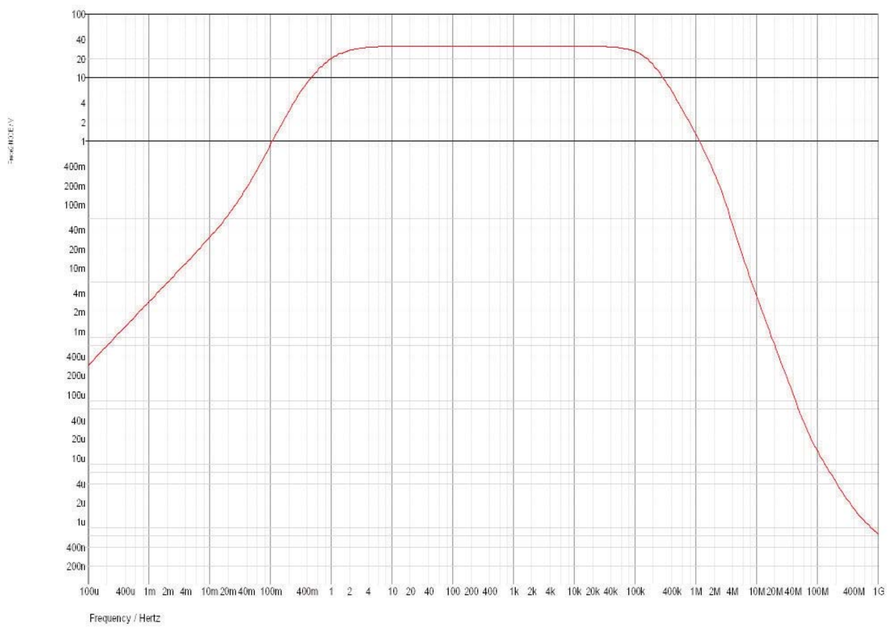


Figure 6: Frequency response of the circuit in Figure 5

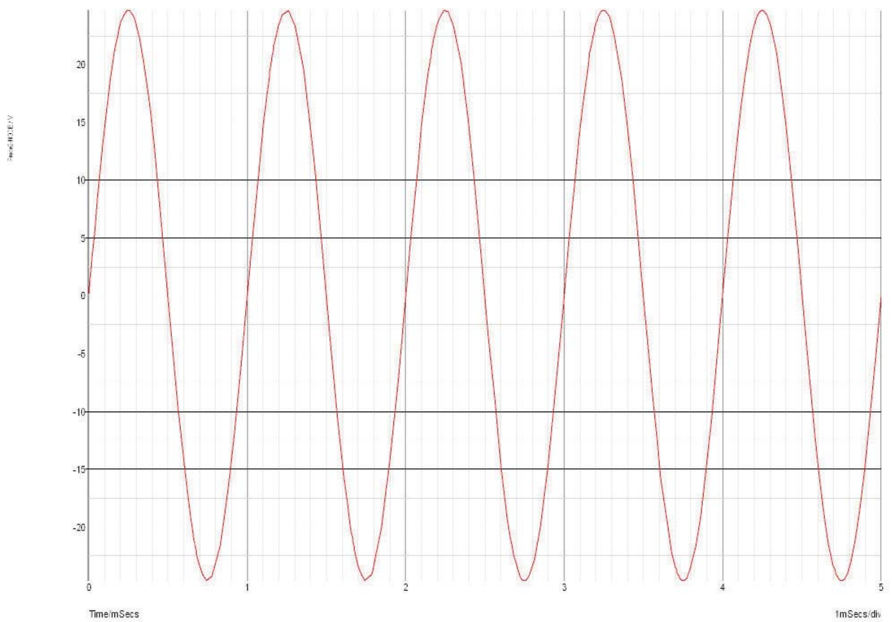


Figure 7: A 1kHz sine waveform amplified by the circuit in Figure 5 (transient distortion 0.05%; 8 Ohms load). It is an extremely pure signal

circuits ruining the sound needs to be investigated further due to their invaluable help in stabilizing the circuit. Thus listening tests are crucial.

Some designers normally put a 220pF capacitor across the bases of the LTP, as shown in **Figure 3b**, my research so far in relation to the topologies I have used shows it does no good and may destabilize the circuit. (Software simulation has been done using SIMetrix/SIMPLIS).

Voltage gain stage

As shown in **Figure 4**, the 10nf capacitor C7 is necessary to increase amp stability so long as C6 or C3 are present. It also aids in the relative phase at high frequency.

It may be possible that transistors don't like to be directly loaded to the collector; when driven from base there isn't a corresponding emitter drive. It may be case that that current mirrors can get away with the loading of their collectors. But, the ears can detect minute variations.

Figure 5 is a fully simulated and tested circuit. Its frequency response is shown in **Figure 6**. Figure **7** shows its behaviour with a 1kHz sine waveform (transient distortion 0.05%; 8 Ohms load). It produces an extremely pure signal. The transient waveform is shown in **Figure 8** at 20kHz.Otherwise it is quite stable and free from oscillation. The purpose of this circuit is to carry out listening tests to establish if it has any listening advantages in terms of reduction in ear fatigue over non-cascode types.

The voltage gain stage normally likes to be protected from the voltage variations that may occur on the supply rail, as the current gain stage is drawing the much needed current for that final speaker drive. Normally the positive rail is dropping (going negative), while the negative rail is climbing (going positive) and the ground/earth is dancing to the beat.

Output bias

Many amplifier developers normally put a capacitor across the bias circuit as shown in **Figure 9** but initial simulations tend to suggest that it does no good to the waveform. It may be claimed that the capacitor is for coupling AC signals across the output bias circuit, but if it is detrimental to the waveform why keep it there.

Protection

A few amplifier designers normally seed the circuit with protection diodes. My view so far is that protection diodes at the input and voltage driver may not be paramount as initially thought, but those at the output may be critical as the safety of the circuit, as without them it may be compromised.

VI limiting is normally essential to keep the transistors within their safe operating

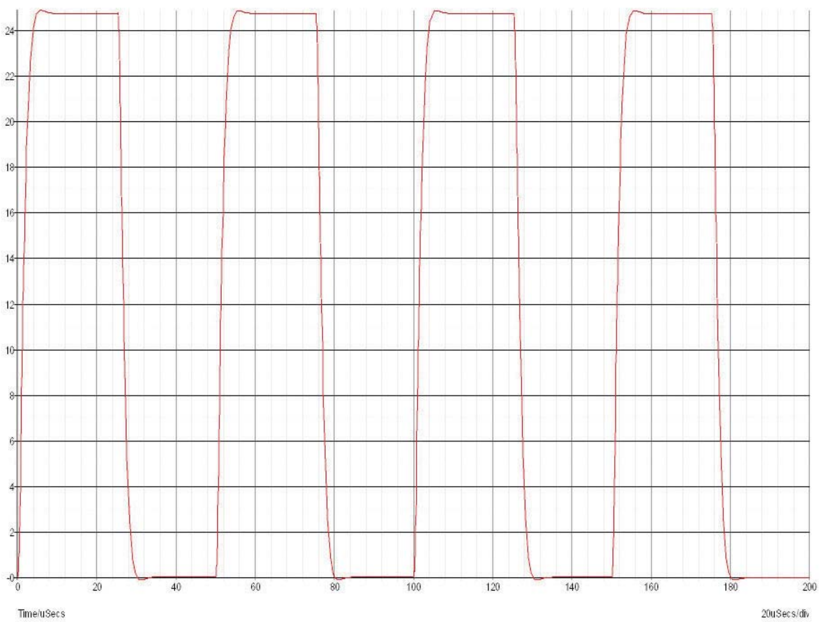


Figure 8: *The amplifier handles transient signals very well as shown in this 20kHz transient waveform. As can be seen, the amplifier is very stable*

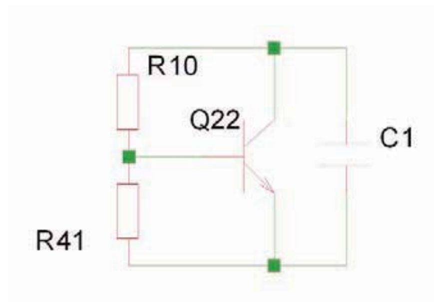


Figure 9: *Putting a capacitor across the bias transistor may degrade the quality of the signal. If a stability capacitor is needed is should be between the base and the collector of the bias transistor*

area and, also, to protect the transistors when the amplifier is connected to a source such as a television, for example. This means that when the television is being switched on, the on and off transistors are not deported to silicon heaven. One thing the VI limiting may not be able to protect against is a dead short.

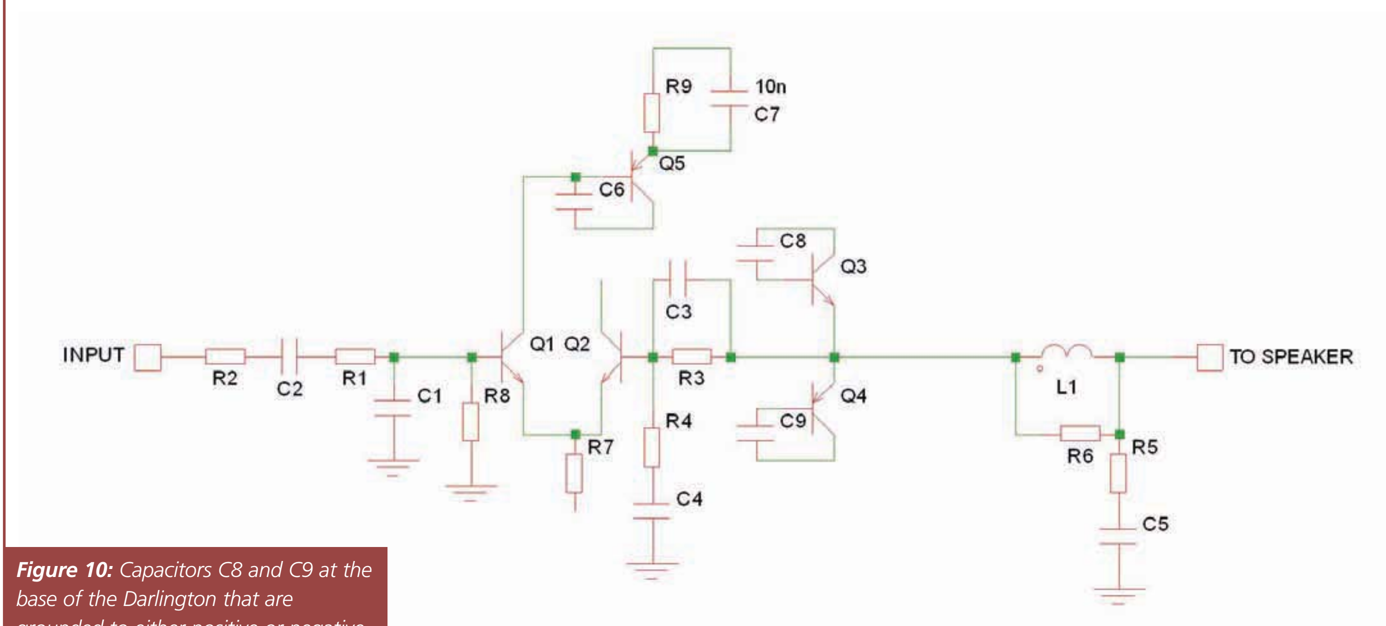


Figure 10: *Capacitors C8 and C9 at the base of the Darlington that are grounded to either positive or negative rail are detrimental according to my research findings and simulation*

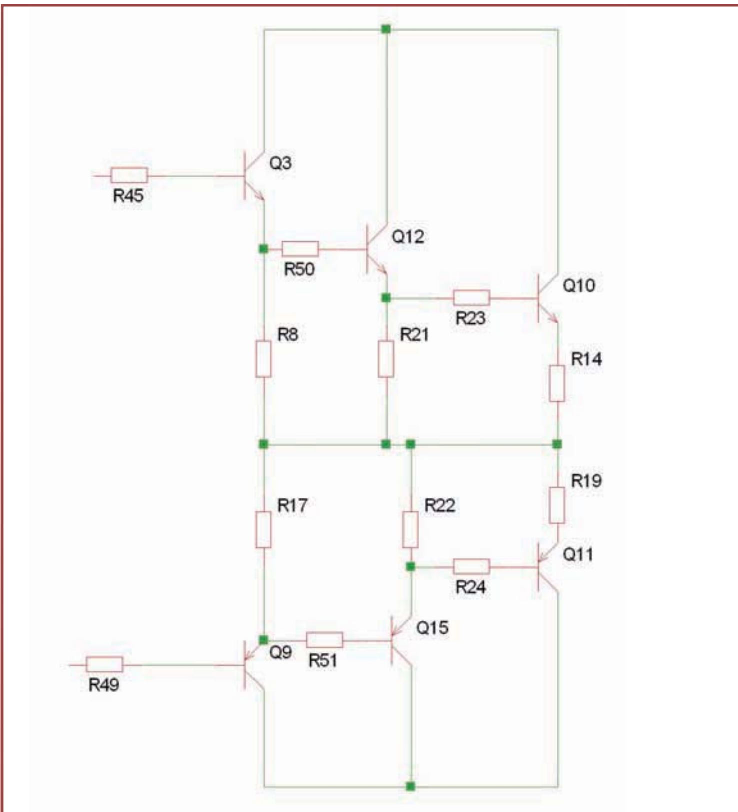


Figure 11: *Resistors R23 and R24 are totally crucial for stability but R50, R51, R45 and R49 spoil the relative phase and destabilize the circuit, though this can be recovered by making sure that the circuit has a high gain*

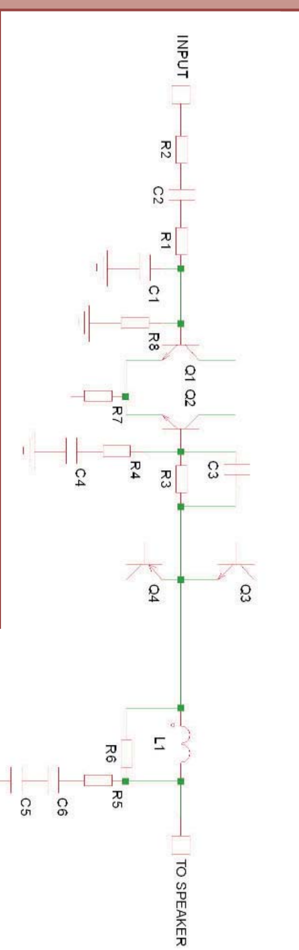


Figure 12: *The output radio frequency snubber R5, C5 and can benefit from having an extra capacitor C6 instead of just one*

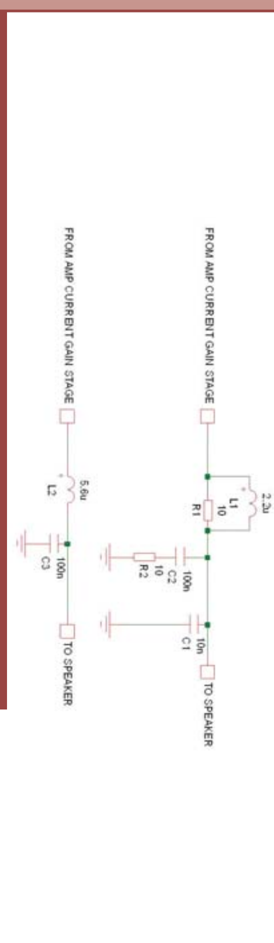


Figure 13: *Techniques for reducing ear fatigue, safeguarding against capacitative loads and safeguarding against radio frequency interference*

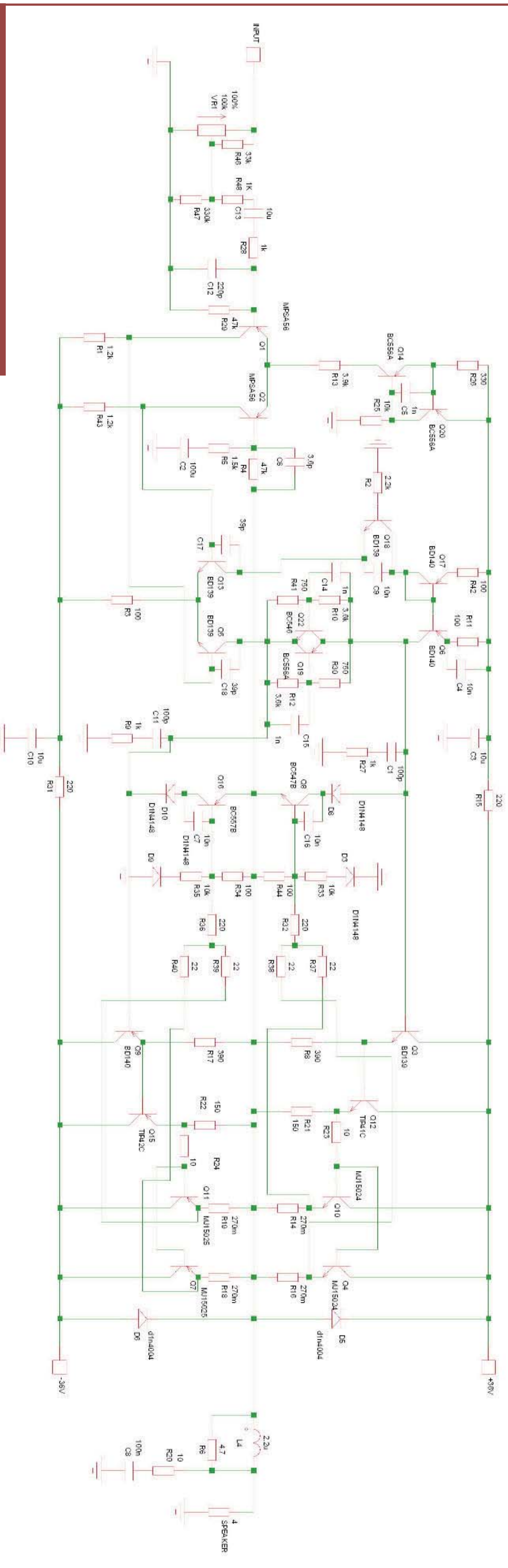


Figure 14: *Sample amplifier for hi-fi listening*

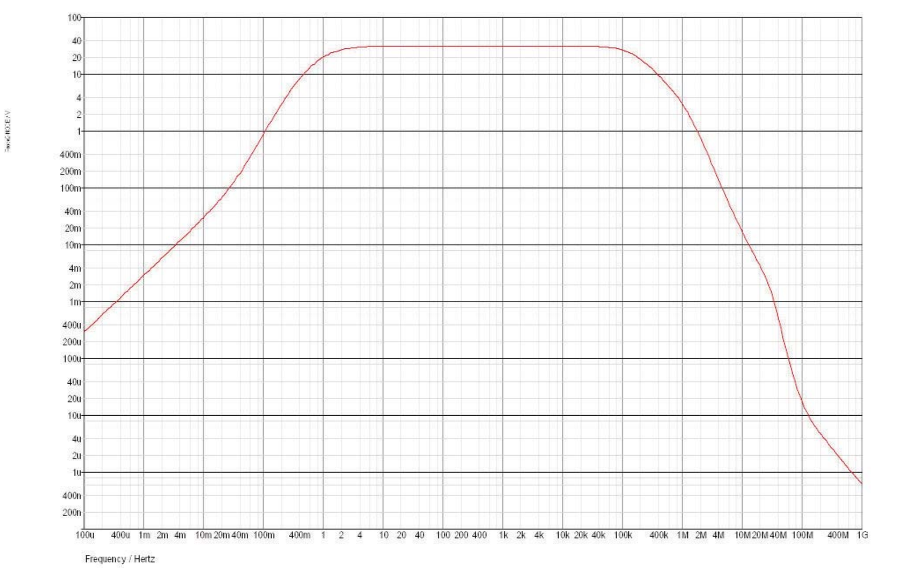


Figure 15: *Amplifier bandwidth*

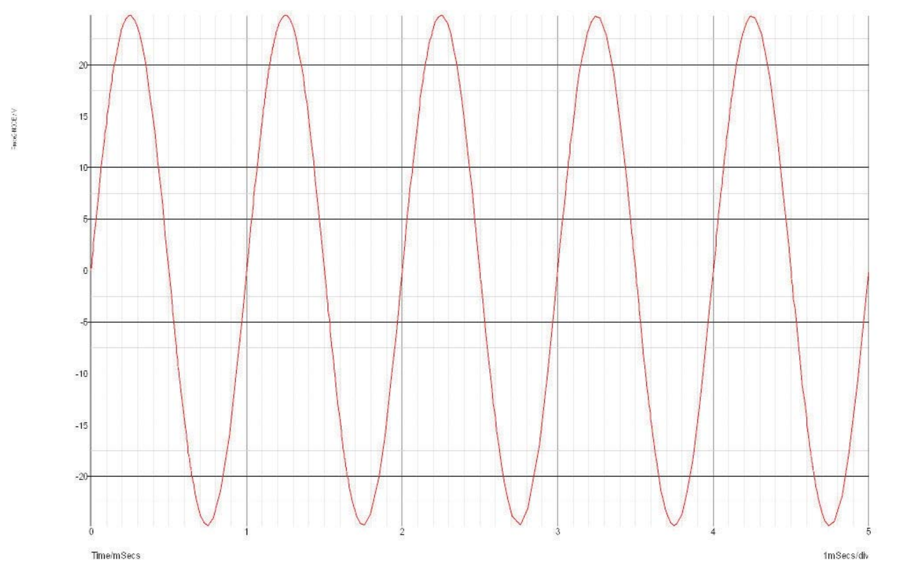


Figure 16: *A 1kHz waveform amplified by the amplifier in Figure 14 with a transient distortion of 0.037%*

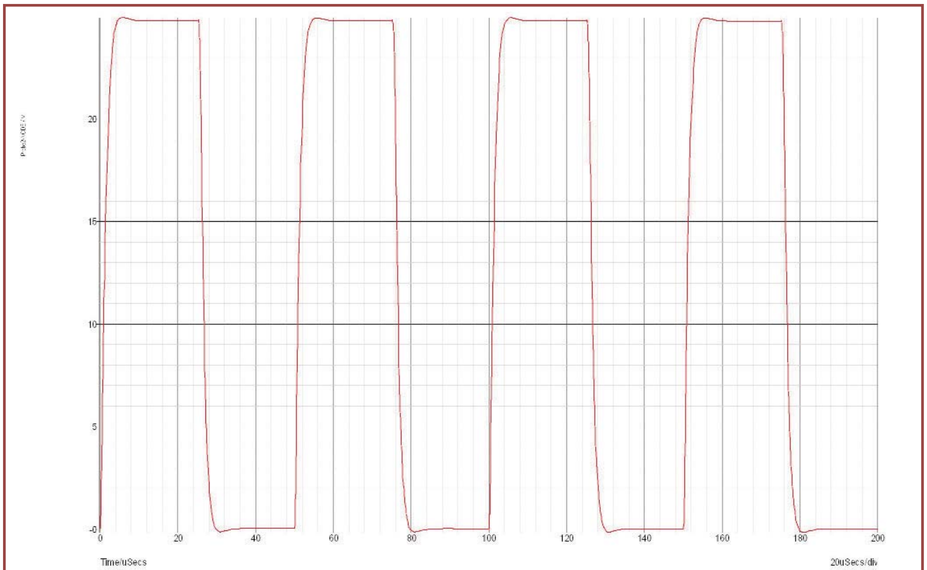


Figure 17: *The amplifier handles transient signals very well, as shown in this 20kHz transient waveform. As can be seen the amplifier is very stable*

Output

There are a few things that some amplifier designers tend to keep doing that makes me wonder, see **Figure 10**. The capacitors (C8 and C9) at the bases of the Darlington that are grounded to either positive or negative rail are detrimental according to my research findings and simulation. Yet a few manufacturers can't seem to do without them.

Then comes the issue of resistors at the bases of transistors. Using **Figure 11** as an example, we have resistors that are totally crucial for stability (R23, R24), then another set that spoils the relative phase (R50, R51) and destabilizes the circuit, and another set that also spoils the relative phase and destabilizes the circuit (R45, R49) (although the stability can be recovered by making sure that the circuit has a high gain). The only reason I can find for them being there is to safeguard the earlier stages from failure in case the output stage blows. Could it be that they may also serve to eliminate ear fatigue?

Then comes **Figure 12**, where we have two capacitors C5 and C6 in series at the output filter. These could serve to increase reliability, whereby in case one of the capacitors was manufactured defective then the other one comes in handy.

Figure 13 shows the techniques that can be used to reduce ear fatigue caused by the speaker. These techniques are used by some manufacturers.

The RL filter at the output of the amplifiers i.e. L1 AND R6 in Figure 12 helps safeguard the amplifier from oscillating while handling capacitative loads.

Sample amplifier

Figure 14 is a sample amplifier for hi-fi listening. **Figure 15** shows the bandwidth, while **Figure 16** shows a 1kHz waveform with a transient distortion of 0.037%. This amplifier has a DC offset of < 1mV. **Figure 17** shows a 20kHz transient.

Conclusion

Class A/B still rocks and may not be replaced by digital too soon – rather the two may coexist. ■

Associate Professor **Dr Murat Uzam** from Nigde University in Turkey presents a series of articles on a project that focuses on a microcontroller-based PLC. This is the fifth article of the series describing the first set of timer macros

PLC with PIC16F648A Micro

IN THIS ARTICLE, the following timer macros are described: TON_8 (ON delay timer) and TOF_8 (OFF delay timer). The remaining timer macros, namely TEP_8 (Extended Pulse timer) and TOS_8 (Oscillator timer) will appear in the next article.

Timer Macros

Timers can be used in a wide range of applications where a time delay function is required based on an input signal. In this article, four different timer functions, namely on delay timer, off delay timer, pulse timer and oscillator timer, are described.

The definition of 8-bit variables to be used for the timer macros and their allocation in BANK 0 of RAM data memory are both shown in **Figures 1a** and **1b** respectively. The status bits of all of these timers are defined as shown in **Figure 2a**.

(a)

```
;--------------- VARIABLE DEFINITIONS -
        CBLOCK 0x44
        TON8_Q,TOF8_Q,TEP8_Q,TOS8_Q
        endc
        CBLOCK 0x48
        TON8
        endc
        CBLOCK 0x50
        TOF8
        endc
        CBLOCK 0x58
        TEP8
        endc
        CBLOCK 0x60
        TOS8
        endc
        CBLOCK 0x68
        TON8_RED,TOF8_RED,TEP8_RED1,TEP8_RED2,TOS8_RED
        endc
;-----------------------------------------
```

(b)

44h	TON8_Q
45h	TOF8_Q
46h	TEP8_Q
47h	TOS8_Q
48h	TON8
49h	TON8+1
4Ah	TON8+2
4Bh	TON8+3
4Ch	TON8+4
4Dh	TON8+5
4Eh	TON8+6
4Fh	TON8+7
50h	TOF8
51h	TOF8+1
52h	TOF8+2
53h	TOF8+3
54h	TOF8+4
55h	TOF8+5
56h	TOF8+6
57h	TOF8+7
58h	TEP 8
59h	TEP8+1
5Ah	TEP 8+2
5Bh	TEP 8+3
5Ch	TEP 8+4
5Dh	TEP 8+5
5Eh	TEP 8+6
5Fh	TEP 8+7
60h	TOS8
61h	TOS8+1
62h	TOS8+2
63h	TOS8+3
64h	TOS8+4
65h	TOS8+5
66h	TOS8+6
67h	TOS8+7
68h	TON8_RED
69h	TOF8_RED
6Ah	TEP8_RED1
6Bh	TEP8_RED2
6Ch	TOS8_RED

BANK 0

***Figure 1: (a)** The definition of 8-bit variables to be used for the timer macros **(b)** Their allocation in BANK 0 of RAM data memory*

All the 8-bit variables defined for timers must be cleared at the beginning of the PLC operation for a proper operation. For this purpose, the macro "init_tmrs" is defined as shown in **Figure 2b**. This macro must be run after the macro "initialize" explained in the follow-up article in the next issue of Electronics World. The file "tmr_mcr_def.inc" (all the files considered in this article including "tmr_mcr_def.inc" can be downloaded from http://host.nigde.edu.tr/muzam/) contains all timer macros defined for UZAM_PLC.

On Delay Timer (TON)

The on-delay timer can be used to delay setting an output true (ON – 1) for a fixed period of time after an input signal becomes true (ON – 1). The symbol and timing diagram of the on-delay timer (TON) are both shown in **Figure 3**.

As the input signal IN goes true (ON – 1), the timing function is started and, therefore, the elapsed time ET starts to increase. When the elapsed time ET reaches the time specified by the preset time input PT, the output Q goes true (ON – 1) and the elapsed time is held. The output Q remains true (ON – 1) until the input signal IN goes false (OFF – 0). If the input signal IN is not true (ON – 1) longer than the delay time specified in PT, the output Q remains false (OFF – 0). The following section explains the implementation of eight of 8-bit on-delay timers for UZAM_PLC.

Macro "TON_8" (8-bit ON Delay Timer)

The macro "TON_8" defines 8 on-delay timers selected with the num = 0, 1…7.

controller – Part 5

(a)

```
;- defining on delay timer outputs -
#define TON8_Q0 TON8_Q,0
#define TON8_Q1 TON8_Q,1
#define TON8_Q2 TON8_Q,2
#define TON8_Q3 TON8_Q,3
#define TON8_Q4 TON8_Q,4
#define TON8_Q5 TON8_Q,5
#define TON8_Q6 TON8_Q,6
#define TON8_Q7 TON8_Q,7

;- defining off delay timer outputs -
#define TOF8_Q0 TOF8_Q,0
#define TOF8_Q1 TOF8_Q,1
#define TOF8_Q2 TOF8_Q,2
#define TOF8_Q3 TOF8_Q,3
#define TOF8_Q4 TOF8_Q,4
#define TOF8_Q5 TOF8_Q,5
#define TOF8_Q6 TOF8_Q,6
#define TOF8_Q7 TOF8_Q,7

;- defining puls timer outputs -----
#define TEP8_Q0 TEP8_Q,0
#define TEP8_Q1 TEP8_Q,1
#define TEP8_Q2 TEP8_Q,2
#define TEP8_Q3 TEP8_Q,3
#define TEP8_Q4 TEP8_Q,4
#define TEP8_Q5 TEP8_Q,5
#define TEP8_Q6 TEP8_Q,6
#define TEP8_Q7 TEP8_Q,7

;- defining osilator timer outputs -
#define TOS8_Q0 TOS8_Q,0
#define TOS8_Q1 TOS8_Q,1
#define TOS8_Q2 TOS8_Q,2
#define TOS8_Q3 TOS8_Q,3
#define TOS8_Q4 TOS8_Q,4
#define TOS8_Q5 TOS8_Q,5
#define TOS8_Q6 TOS8_Q,6
#define TOS8_Q7 TOS8_Q,7
```

(b)

```
;------------------- macro: init_tmrs ---
init_tmrs   macro
      clrf TON8          ;Clear TON8
      clrf TON8+1        ;Clear TON8+1
      clrf TON8+2        ;Clear TON8+2
      clrf TON8+3        ;Clear TON8+3
      clrf TON8+4        ;Clear TON8+4
      clrf TON8+5        ;Clear TON8+5
      clrf TON8+6        ;Clear TON8+6
      clrf TON8+7        ;Clear TON8+7
      clrf TOF8          ;Clear TOF8
      clrf TOF8+1        ;Clear TOF8+1
      clrf TOF8+2        ;Clear TOF8+2
      clrf TOF8+3        ;Clear TOF8+3
      clrf TOF8+4        ;Clear TOF8+4
      clrf TOF8+5        ;Clear TOF8+5
      clrf TOF8+6        ;Clear TOF8+6
      clrf TOF8+7        ;Clear TOF8+7
      clrf TEP8          ;Clear TEP8
      clrf TEP8+1        ;Clear TEP8+1
      clrf TEP8+2        ;Clear TEP8+2
      clrf TEP8+3        ;Clear TEP8+3
      clrf TEP8+4        ;Clear TEP8+4
      clrf TEP8+5        ;Clear TEP8+5
      clrf TEP8+6        ;Clear TEP8+6
      clrf TEP8+7        ;Clear TEP8+7
      clrf TOS8          ;Clear TOS8
      clrf TOS8+1        ;Clear TOS8+1
      clrf TOS8+2        ;Clear TOS8+2
      clrf TOS8+3        ;Clear TOS8+3
      clrf TOS8+4        ;Clear TOS8+4
      clrf TOS8+5        ;Clear TOS8+5
      clrf TOS8+6        ;Clear TOS8+6
      clrf TOS8+7        ;Clear TOS8+7
      clrf TON8_Q        ;Clear TON8_Q
      clrf TOF8_Q        ;Clear TOF8_Q
      clrf TEP8_Q        ;Clear TEP8_Q
      clrf TOS8_Q        ;Clear TOS8_Q
      clrf TON8_RED      ;Clear TON8_RED
      clrf TOF8_RED      ;Clear TOF8_RED
      clrf TEP8_RED1     ;Clear TEP8_RED1
      clrf TEP8_RED2     ;Clear TEP8_RED2
      clrf TOS8_RED      ;Clear TOS8_RED
      endm
;--------------------------------------
```

***Figure 2: (a)** The definition of status bits of timer macros*
***(b)** The initialization of all variables of timer macros defined as a macro "init_tmrs"*

Table 1 shows the macro "TON_8" and its symbol. IN (input signal), Q (output signal = timer status bit) and CLK {free running timing signals – ticks: T00(0.512ms)…T15(16777.216ms)} are all defined as Boolean variables.

The time constant "tcnst" is an integer constant (here for 8-bit resolution, it is chosen as any number in the range 1-255) and is used to define preset time PT, which is obtained by the formula: PT = tcnst × CLK, where CLK should be used as the period of the free-running timing signals – ticks.

The on-delay timer outputs are represented by the status bits: TON8_Q,num (num = 0, 1…7), namely TON8_Q0, TON8_Q1…TON8_Q7, as shown in Figure 2a. We use a Boolean variable, or TON8_RED,num (num = 0, 1…7), as a rising edge detector for identifying the rising edges of the chosen CLK. An 8-bit integer variable TON8+num (num = 0, 1…7) is used to count the rising edges of the CLK. The count value of TON8+num (num = 0, 1…7) defines the elapsed time ET as follows: ET = CLK × count value of TON8+num (either of 0, 1…7).

Let us now briefly consider how the macro "TON_8" works. First of all, preset time PT, is defined by means of a reference timing signal "CLK = t_reg,t_bit" and a time constant "tcnst". If the input signal IN, taken into the macro by means of W is false (OFF – 0), then the output signal TON8_Q,num (num = 0, 1…7) is forced to be false (OFF – 0) and the counter TON8+num (num = 0, 1…7) is loaded with "00h".

If the input signal IN is true (ON – 1) and the output signal Q, i.e. the status bit

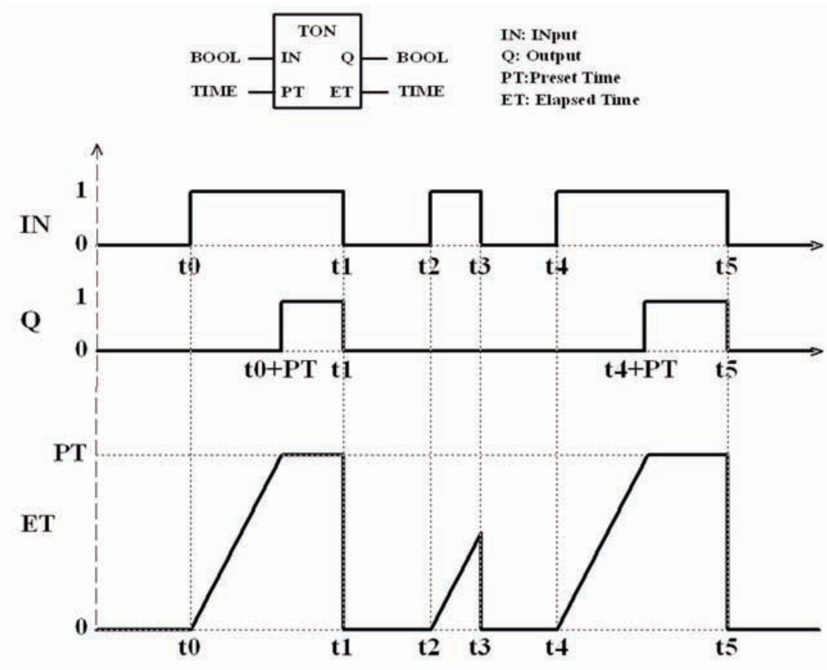
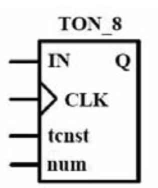


Figure 3: *The symbol and timing diagram of the on-delay timer (TON)*

Macro

```
;--------- macro: TON_8 ------------
TON_8 macro     num,t_reg,t_bit,tcnst
      local     L1,L2
      movwf     Temp_1
      btfsc     Temp_1,0
      goto      L2
      movlw     00h
      movwf     TON8+num
      bcf       TON8_Q,num
      goto      L1
L2    btfsc     TON8_Q,num
      goto      L1
      btfss     t_reg,t_bit
      bsf       TON8_RED,num
      btfss     t_reg,t_bit
      goto      L1
      btfss     TON8_RED,num
      goto      L1
      bcf       TON8_RED,num
      incf      TON8+num,f
      movfw     TON8+num
      xorlw     tcnst
      skpnz
      bsf       TON8_Q,num
L1
      endm
;----------------------------------
```

Symbol



PT = tcnst x CLK

IN : (W) = 0, 1

CLK (t_reg,t_bit) = T00(0.512 ms), …, T15(16777.216 ms)

tcnst (8bit) = 1, 2, ..., 255

num = 0, 1, …, 7

Q = TON8_Q,num (num = 0, 1, …, 7)

Table 1: *The macro "TON_8" and its symbol*

TON8_Q,num (num = 0, 1…7), is false (OFF – 0), then with each "rising edge" of the reference timing signal "CLK = t_reg,t_bit" the related counter "TON8+num" is incremented by one. In this case, when the count value of "TON8+num" is equal to the number "tcnst", then state-change from 0 to 1 is issued for the output signal (timer status bit) TON8_Q,num (num = 0, 1…7).

If the input signal IN and the output signal Q, i.e. the status bit TON8_Q,num (num = 0, 1…7) are both true (ON – 1), then no action is taken and the elapsed time ET is held. In this macro a previously defined 8-bit variable "Temp_1" is also utilized.

Off Delay Timer (TOF)

The off-delay timer can be used to delay setting an output false (OFF – 0) for a fixed period of time after an input signal goes false (OFF – 0), i.e. the output is held ON for a given period longer than the input. The symbol and timing diagram of the off-delay timer (TOF) are both shown in **Figure 4**.

As the input signal IN goes true (ON – 1), the output Q follows and remains true (ON – 1), until the input signal IN is false (OFF – 0) for the period specified in preset time input PT. As the input signal IN goes false (OFF – 0), the elapsed time ET starts to increase. It continues to increase until it reaches the preset time input PT, at which point the output Q is set false (OFF – 0) and the elapsed time is held. If the input signal IN is only false (OFF – 0) for a period shorter than the input PT, the output Q remains true (ON – 1). The following section explains the implementation of eight of 8-bit off-delay timers for UZAM_PLC.

Macro "TOF_8" (8-bit OFF Delay Timer)

The macro "TOF_8" defines 8 off-delay timers selected with the num = 0, 1…7. **Table 2** shows the macro "TOF_8" and its symbol. IN (input signal), Q (output signal

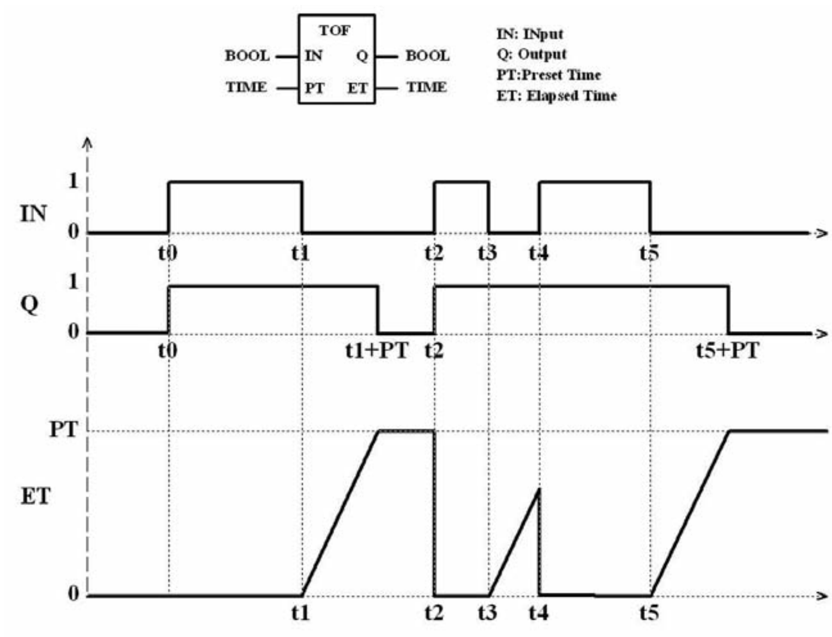
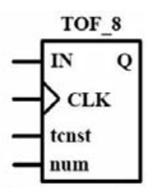


***Figure 4:** The symbol and timing diagram of the off-delay timer (TOF)*

Macro	***Symbol***

```
;--------- macro: TOF_8 ------------
TOF_8 macro   num,t_reg,t_bit,tcnst
      local   L1,L2
      movwf   Temp_1
      btfss   Temp_1,0
      goto    L2
      movlw   00h
      movwf   TOF8+num
      bsf     TOF8_Q,num
      goto    L1
L2    btfss   TOF8_Q,num
      goto    L1
      btfss   t_reg,t_bit
      bsf     TOF8_RED,num
      btfss   t_reg,t_bit
      goto    L1
      btfss   TOF8_RED,num
      goto    L1
      bcf     TOF8_RED,num
      incf    TOF8+num,f
      movfw   TOF8+num
      xorlw   tcnst
      skpnz
      bcf     TOF8_Q,num
L1
      endm
;----------------------------------
```



PT = tcnst x CLK

IN = 0, 1

CLK (t_reg,t_bit) = T00(0.512 ms), …, T15(16777.216 ms)

tcnst (8bit) = 1, 2, ..., 255

num = 0, 1, …, 7

Q = TOF8_Q,num (num = 0, 1, …, 7)

***Table 2:** The macro "TON_8" and its symbol*

= timer status bit) and CLK {free running timing signals – ticks: T00(0.512ms)…T15(16777.216ms)} are all defined as Boolean variables.

The time constant "tcnst" is an integer constant (here for 8-bit resolution it is chosen as any number in the range 1-255) and is used to define preset time PT, which is obtained by the formula: PT = tcnst × CLK, where CLK should be used as the period of the free-running timing signals – ticks. The off-delay timer outputs are represented by the status bits: TOF8_Q,num (num = 0, 1…7), namely TOF8_Q0, TOF8_Q1…TOF8_Q7, as shown in Figure 2a.

We use a Boolean variable, namely TOF8_RED,num (num = 0, 1…7), as a rising edge detector for identifying the rising edges of the chosen CLK. An 8-bit integer variable TOF8+num (num = 0, 1…7) is used to count the rising edges of the CLK. The count value of TOF8+num (num = 0, 1…7) defines the elapsed time ET as follows: ET = CLK × count value of TOF8+num (either of 0, 1…7).

Let us now briefly consider how the macro "TOF_8" works. First of all, preset time PT, is defined by means of a reference timing signal "CLK = t_reg,t_bit" and a time constant "tcnst". If the input signal IN, taken into the macro by means of W, is true (ON – 1), then the output signal TOF8_Q,num (num = 0, 1…7) is forced to be true (ON – 1) and the counter TOF8+num (num = 0, 1…7) is loaded with "00h".

When IN = 1 and TOF8_Q,num = 1, if IN goes false (OFF – 0), then with each "rising edge" of the reference timing signal "CLK = t_reg,t_bit" the related counter "TOF8+num" is incremented by one. In this case, when the count value of "TOF8+num" is equal to the number "tcnst", then a state-change from 1 to 0 is issued for the output signal (timer status bit) TOF8_Q,num (num = 0, 1…7). In this macro a previously defined 8-bit variable "Temp_1" is also utilized. ■

Special DATA Acquisition Board

Maurizio Di Paolo Emilio presents a series of articles focusing on a data acqusition (DAQ) board project for the management of environmental sensors and a high-speed data acquisition system. This article, the second in the series, links the board to a GPS and ADC.

A DATA ACQUISITION (DAQ) board for managing environmental sensors and a high speed data acquisition system is discussed here. It is composed of various slots for the upgrade. It can work on a VME bus but also by the means of a USB, GSM, Wireless and Ethernet connection.

It comprises the following parts: a GPS, an analogue-to-digital converter (ADC), an FPGA, a serial port and a scaler. In particular we want to analyze how to get the time and acquire the data. The goal is to have a time reference for each data that the board acquires.

The Time: Global Positioning System

The Global Positioning System (GPD) consists of three interacting components:

1) The space segment – satellites orbiting the earth;
2) The control segment – the control and monitoring stations run by the United States' Department of Defense; and
3) The user segment – the GPS receivers owned by civilians and the military.

The space segment consists of a constellation of 24 active satellites (and one or more in-orbit spares), orbiting the earth every 12 hours. Four satellites are located in each of the six orbits and will be visible from any location on each, 95% of the time. The orbits are distributed evenly around the earth and are inclined 55 degrees from the equator. The satellites orbit at an altitude of about 11,000 nautical miles.

Each satellite transmits two signals: L1 (1575.42MHz) and L2 (1227.60MHz). The L1 signal is modulated with two pseudo-random noise signals – the protected (P) code and the course/acquisition (C/A) code. The L2 signal only carries the P code. Civilian navigation receivers only use the C/A code on the L1 frequency.

Each signal from each satellite contains a repeating message, indicating the position and orbital parameters of itself and the other satellites (almanac), a bill of health for the satellites (health bit) and the precise atomic time.

The receiver used for this application is that of Parallax (**Figure 1**). The receiver measures the time required for the signal to travel from the satellite to the receiver, by knowing the time when the signal left the satellite and observing the time it receives the signal, based on its internal clock.

If the receiver had a perfect clock, exactly in sync with those on the satellites, three measurements from three satellites would be sufficient to determine position in three dimensions via triangulation. However, that is not the case, so a fourth satellite is needed to resolve the receiver clock error. With four satellites, a GPS receiver can provide very accurate clock (time, date) and position information (latitude, longitude, altitude, speed, travel direction/heading).

The GPS receiver module (**Figures 2** and **3**) provides standard, raw NMEA0183 (National Marine Electronics Association) strings or specific user-requested data via the serial command interface, tracking of up to 12 satellites, and WAAS/EGNOS (Wide Area Augmentation System/European Geostationary Navigation Overlay Service) functionality for more accurate positioning results.

The module provides current time, date, latitude, longitude,

Pin	Pin Name	Type	Function
1	GND	G	System ground. Connect to power supply's ground (GND) terminal.
2	VCC	P	System power, +5V DC input.
3	SIO	I/O	Serial communication (commands sent TO the Module and data received FROM the Module). Asynchronous, TTL-level interface, 4800bps, 8 data bits, no parity, 1 stop bit, non-inverted.
4	/RAW	I	Mode select pin. Active LOW digital input. Internally pulled HIGH by default. When the /RAW pin is unconnected, the default "Smart Mode" is enabled, wherein commands for specific GPS data can be requested and the results will be returned (see the "Command Structure" section for more details). When /RAW is pulled LOW, the Module will enter "Raw Mode" and will transmit standard strings, allowing advanced users to use the raw GPS data directly.

Note: Type: I = Input, O = Output, P = Power, G = Ground

***Figure 2:** Electronic outline of GPS*

PART 2

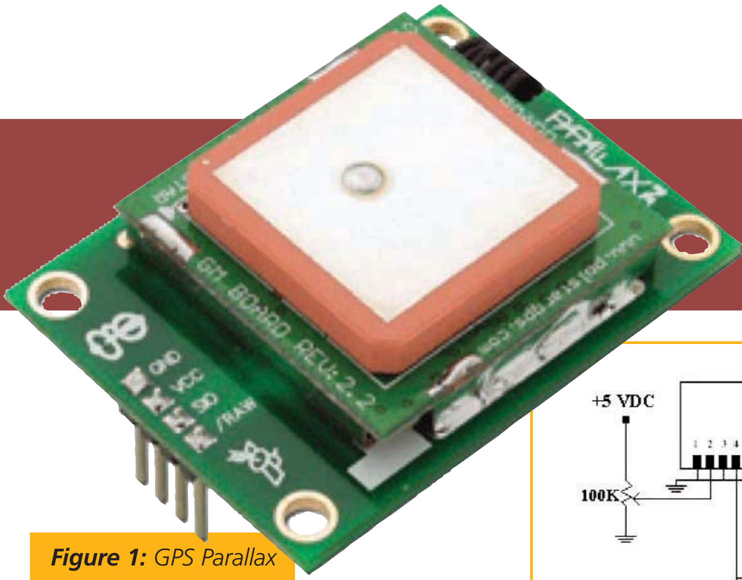


Figure 1: *GPS Parallax*

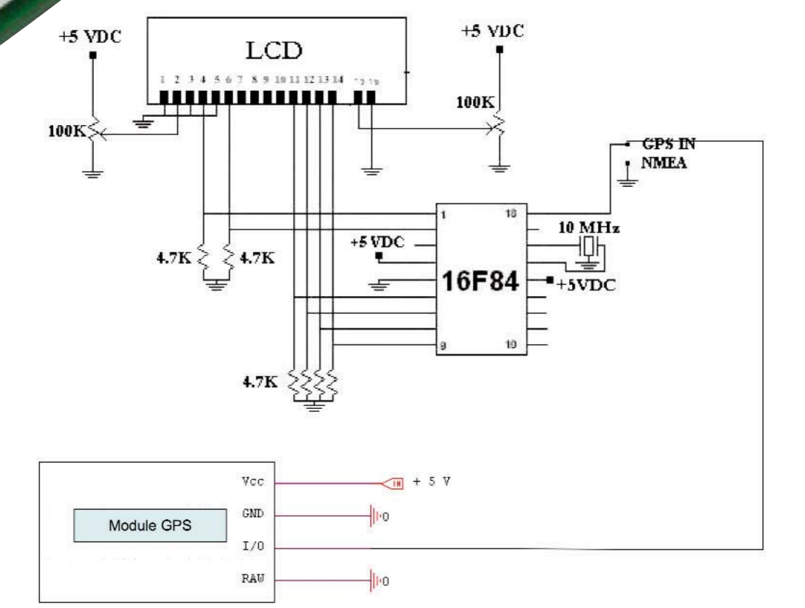


Figure 3: *PIN of GPS receiver*

altitude, speed and travel direction/heading among other data, and can be used in a wide variety of hobbyist and commercial applications, including navigation, tracking systems, mapping, fleet management, auto-pilot and robotics.

Module Highlights

- Fully-integrated, low-cost GPS receiver module with on-board, passive patch antenna;
- Single-wire, 4800 baud Serial TTL interface to BASIC Stamp SX, Propeller and other processors;
- Provides either raw NMEA0183 strings or specific data requested via the command interface;
- Requires single +5VDC supply @ 115mA (typical);
- 0.100" pin spacing for easy prototyping and integration;
- Programmable Parallax SX/B microprocessor and open-source control firmware for advanced users (not supported by Parallax, but offered as a download from the Parallax web site).

Most GPS receivers output a stream of data so that it can be used and interpreted by other devices. The most common format (and used by our GPS Receiver Module in "Raw Mode") is NMEA0183 (National Marine Electronics Association, http://www.nmea.org/), developed for data communications between marine instruments. Some receivers also have proprietary data formats which are used (in the case of navigation receivers) to transfer waypoint lists, track logs and other data between the GPS and a computer. Such proprietary formats are not covered by the NMEA standard.

NMEA 0183 (or NMEA for short) is a combined electrical and data specification for communication between marine electronic devices such as echo sounder, sonars, anemometer (wind speed and direction), gyrocompass, autopilot, GPS receivers and many other types of instruments. It has been defined and controlled by the US-based National Marine Electronics Association.

The NMEA 0183 standard uses a simple ASCII, serial communications protocol that defines how data is transmitted in a "sentence" from one "talker" to one "listener" at a time. Through the use of intermediate expanders, a talker can have a unidirectional conversation with multiple listeners and using multiplexers, multiple sensors can talk to a single computer port. Third-party switches are available that can establish a primary and secondary talker, with automatic failover if the primary fails.

Analogue to Digital Converter

An analogue-to-digital converter (abbreviated ADC, A/D or A to D) is a device that converts continuous signals to discrete digital numbers. The reverse operation is performed by a digital-to-analogue converter (DAC).

Typically, an ADC is an electronic device that converts an input analogue voltage (or current) to a digital number. However, some non-electronic or only partially electronic devices, such as rotary encoders, can also be considered ADCs. The digital output may use different coding schemes, such as binary, Gray code or two's complement binary.

Most converters sample with 6 to 24 bits of resolution and produce fewer than one megasample per second. Thermal noise generated by passive components such as resistors masks the measurement when higher resolution is desired. For audio applications and in room temperatures, such noise is usually a little less than 1μV (microvolt) of white noise. If the Most Significant Bit (MSB) corresponds to a standard 2V of output signal, this translates to a noise-limited performance that is less than 20~21 bits and obviates the need for any dithering.

Mega- and gigasample per second converters are also available. Megasample converters are required in digital video cameras, video

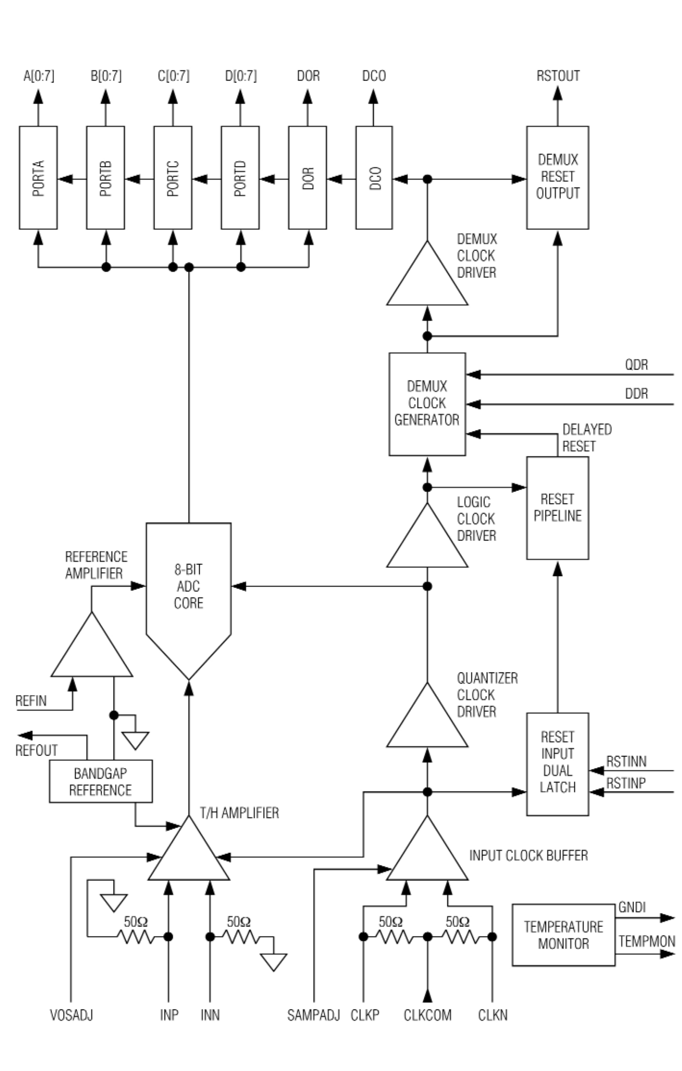


Figure 4: *MAX109*

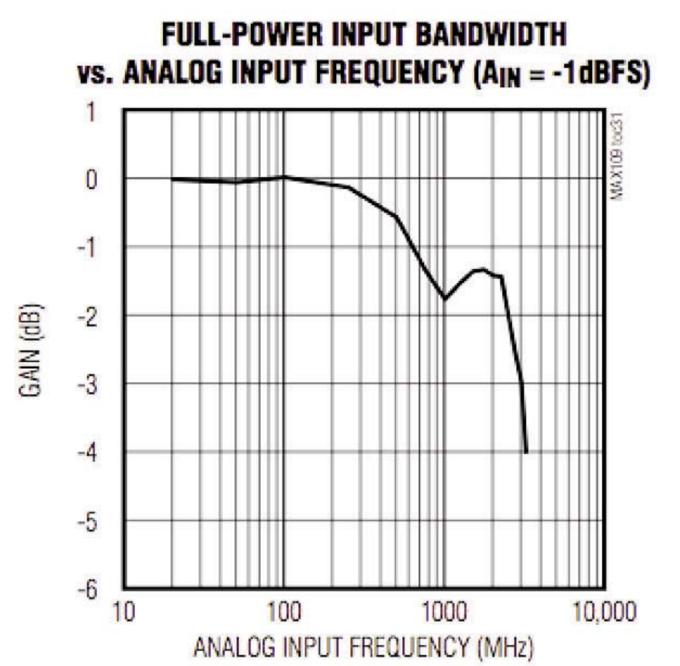


Figure 5: *Characteristics of MAX109*

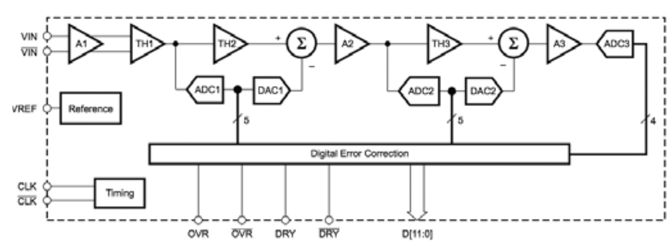


Figure 6: *Outline ADS 5463*

capture cards and TV tuner cards to convert full-speed analogue video to digital video files. Commercial converters usually have ±0.5 to ±1.5 LSB (Least Significant Bit) error in their output.

In many cases the most expensive part of an integrated circuit is the pins, because they make the package larger and each pin has to be connected to the integrated circuit's silicon. To save pins, it's common for slow ADCs to send their data one bit at a time over a serial interface to the computer, with the next bit coming out when a clock signal changes state, say from zero to 5V. This saves quite a few pins on the ADC package and, in many cases, does not make the overall design any more complex. (Even microprocessors which use memory-mapped IO only need a few bits of a port to implement a serial bus to an ADC.)

Commercial ADCs often have several inputs that feed the same converter, usually through an analogue multiplexer. Different models of ADC may include sample-and-hold circuits, instrumentation amplifiers or differential inputs, where the quantity measured is the difference between two voltages.

Implementing ADCs

These are the most common ways of implementing an electronic ADC:

- Direct conversion ADC or flash ADC.
- Successive-approximation ADC.
- Ramp-compare ADC (also called integrating, dual-slope or multi-slope ADC); produces a saw-tooth signal that ramps up, then quickly falls to zero.
- Delta-encoded ADC has an up-down counter that feeds a digital-to-analogue converter (DAC).
- Pipeline ADC (also called subranging quantizer) uses two or more steps of subranging.
- Sigma-Delta ADC (also known as a Delta-Sigma ADC) oversamples the desired signal by a large factor and filters the desired signal band.

There can be other ADCs that use a combination of electronics and other technologies:

- Time-stretch analogue-to-digital converter (TS-ADC); digitizes a very wide bandwidth analogue signal that cannot be digitized by a conventional electronic ADC, by time-stretching the signal prior to digitization.

In the DAQ board we use three types of ADCs:

- **ADC MAX109**

The MAX109, 2.2Gsps, 8-bit, analogue-to-digital converter

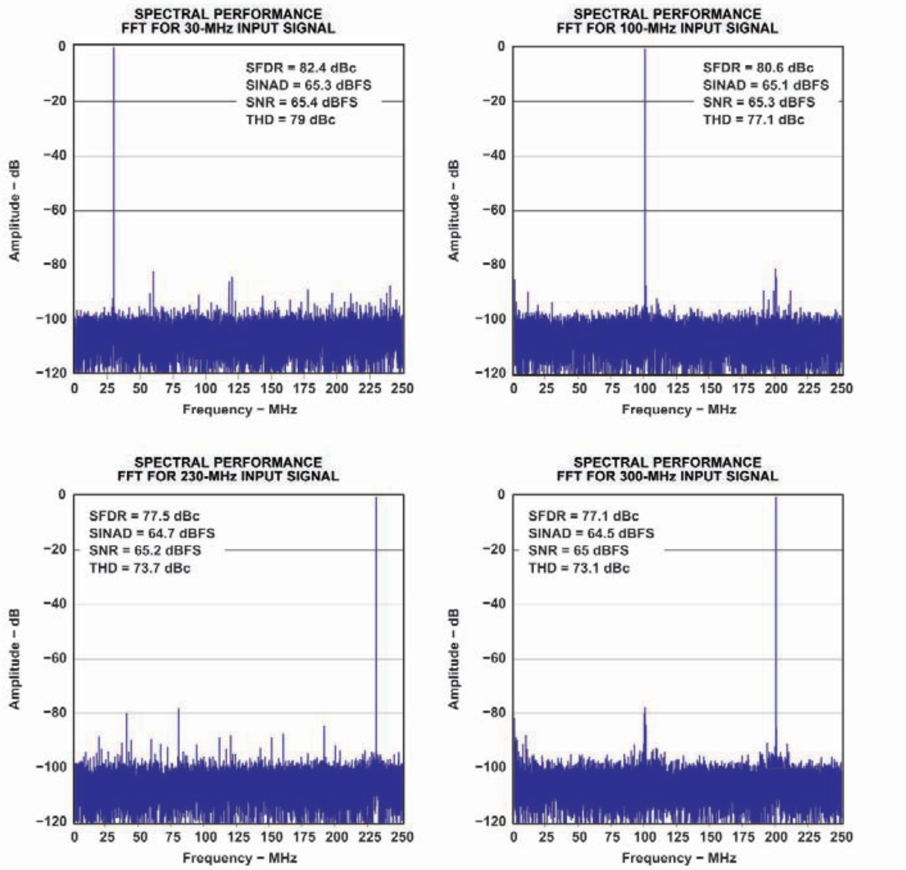


Figure 7: *Data sheet ADS 5463: Spectrum*

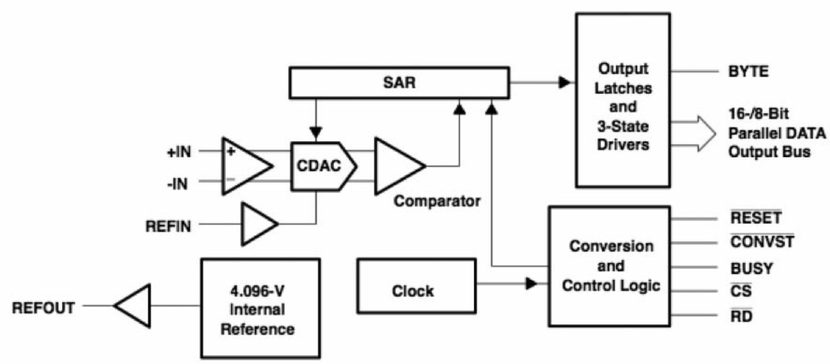


Figure 8: *ADS 5463*

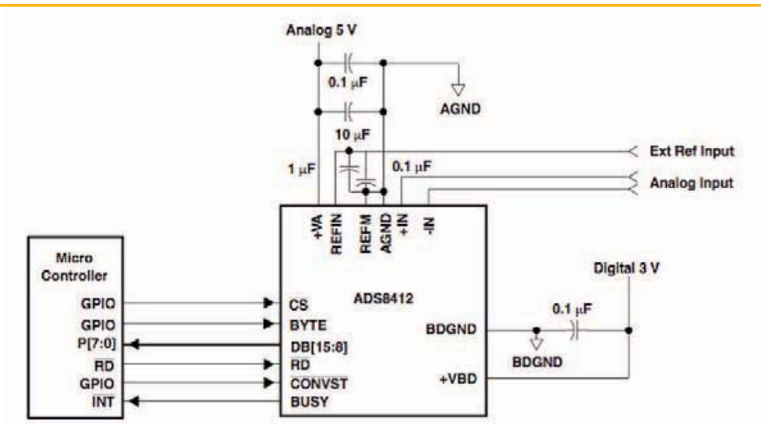


Figure 9: *Application of ADS 5463: Microcontroller interfacing*

(ADC) enables the accurate digitizing of analogue signals with frequencies up to 2.5GHz ultra-high speed, 8-bit, 2.2Gsps ADC.

The innovative design of the internal T/H amplifier, which has a wide 2.8GHz full-power bandwidth, enables a flat-frequency response through the second Nyquist region. This results in excellent ENOB performance of 6.9 bits.

ADS 5463

The ADS5463 is a 12-bit, 500MSPS analogue-to-digital converter (ADC) that operates from both a 5V supply and 3.3V supply, while providing LVDS-compatible digital outputs. This ADC is one of a family of 12, 13 and 14-bit ADCs that operate from 210-500MSPS. The ADS5463 input buffer isolates the internal switching of the onboard track and hold (T&H) from disturbing the signal source, while providing a high-impedance input. An internal reference generator is also provided to simplify the system design.

Designed with a 2.3GHz input bandwidth for the conversion of signals that exceed 500MHz of input center frequency at 500MSPS, the ADS5463 has outstanding low noise performance and spurious-free dynamic range over a large input frequency range.

ADS 8412

The ADS8412 is a 16-bit, 2MHz A/D converter with an internal 4.096-Vref. The device includes a 16-bit, capacitor-based, SARA/D converter with inherent sample-and-hold. The ADS8412 offers a full 16-bit interface and an 8-bit option where data is read using two 8-bit read cycles.

The ADS8412 has a unipolar differential input. It is available in a 48-lead TQFP package and is characterized over the industrial -40°C to 85°C.

The series continues in the next issue of Electronics World magazine. If you missed your last copy you can order your digital version of that issue on line at www.electronicsworld.co.uk

WIN AN LED-DRIVING CONSTANT CURRENT DEMO BOARD!

Electronics World is offering its readers the chance to win a new demonstration board for LED applications from V•I Chip, Inc., a subsidiary of Vicor Corporation. The constant current PRM™ regulator board demonstrates the precisely regulated current output of the high-power density V•I Chip PRM. These devices deliver 99.7% across the load range, meeting the demands of direct-drive multi-LED applications where the intensity and brightness are controlled by regulating the current through the LEDs. The board can be used as a standalone non-isolated source to provide adjustable current up to 240 W (5 A at 48 V), or combined with V•I Chip VTM™ transformers to provide an adjustable isolated current up to 100 A. Paired PRM+VTM modules use less than 1 W for every 1,000 Lumens generated by the LEDs, enabling high-efficiency applications. This is a perfect complement to using BCM™ bus converters with low-voltage driver ICs for lower power applications such as LED TV backlighting.

The V•I Chip regulator board has Kelvin connections for measuring the efficiency of the V•I Chip components, independent of load connect losses, whilst oscilloscope probe jacks are available for measuring output voltage, including output voltage ripple. The board has fused PRM inputs and provision for mounting an optional V•I Chip pushpin heat sink, as well as system enable and disable.

To enter the prize draw send an email to surope@vicorpower.com, with the subject 'LED Competition' with a description of an application that could make use of the V•I Chip PRM driving LEDs. Your description should be less than 50 words. The winner will be both innovative and make best use of the benefits of the PRM. All entries must be received by April 30th 2009 and the winners name will be published in a subsequent issue of Electronics World.

To enter the prize draw send an email to:

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with the subject 'LED Competition' with a description of an application that could make use of the V•I Chip PRM driving LEDs. Your description should be less than 50 words. The winner will be both innovative and make best use of the benefits of the PRM. All entries must be received by April 30th 2009 and the winners name will be published in a subsequent issue of Electronics World.

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In conjunction with Electronics World, Lantronix is offering readers a chance to win one of the SecureLinx Spider™ next generation KVM-over-IP kits, worth over £300. The kit provides secure KVM (keyboard, video, mouse) server management over an IP network. Unlike any other product on the market, Spider offers a flexible, scalable and affordable CAT5-based remote access KVM solution in a cable friendly, compact "zero-U" footprint package. Connected directly to the server, Spider guarantees non-blocked access from any web browser, anywhere, at any time! More cost-effective than traditional KVM, it provides one of the lowest "cost-per-remote-user" server management solutions available. Furthermore no client software or external power supply is required.

For a chance to win a copy of **Lantronix SecureLinx Spider kit**, please answer the following question: **WHAT DOES KVM STAND FOR?** Please send you response to:

lantronix.competition@livewirepr.com

with the following information: Name • E-mail • Job title (student if applicable) • Company/Educational institution • Address (Work or Home to be specified) • Tel No.

Please mark the subject heading as "Electronics World competition entry"

FIT FOR PURPOSE FROM COMMODITY TO NICHE APPLICATIONS

By Chris Williams, UKDL

ON MY OFFICE desk sit two LCD monitors, each driven by a separate computer. One is a 24” 1600x1200 (2M pixels) display, the other is a 30” 2560x1600 (4M pixels) display. Is this necessity or an indulgence?

The answer, as always, is: it depends on the application. Usability of a display and the number of pixels needed to perform a task safely and robustly depends on:

- Ambient lighting levels (lux)
- Positions of light sources and reflections
- Task to be performed
- The individual using the equipment
- Display brightness, resolution and contrast.

Displays used outdoors are at the mercy of sunlight, which changes with orientation, latitude, time of day, time of year and the weather. In addition, ambient lighting can give rise to specular reflection and glare, which affects the perceived contrast.

The most stringent requirements and specifications I have seen for outdoor electronic displays are for cockpit displays in fighter aircraft, which are subjected to extremes of lighting, temperature, shock and vibration during their operation. Most other displays have a more forgiving operating environment and for other indoor mission-critical applications the environment can often be modified to maximise the efficacy of the display parameters and to ensure that the operator who has to read data from a display can do so quickly, safely and without unnecessary distraction.

Tasks such as air traffic control and medical analysis are often carried out in rooms with controlled, low-level lighting. At the present time only liquid crystal displays are available as colour display monitors for such high-resolution applications. As explained in a previous article, light from the backlight of an LCD is absorbed by the polarisers and colour filters, and only ~ 5-7% of the light that impinges on the rear of the LCD display will eventually emerge from the front of the display to be read by an operator. In mission-critical applications such as medical X-ray diagnosis, where the misreading or a shadow within a medical image could result in misdiagnosis of a patient, it is essential that the ambient lighting is adjusted to avoid impairing the efficacy of the display.

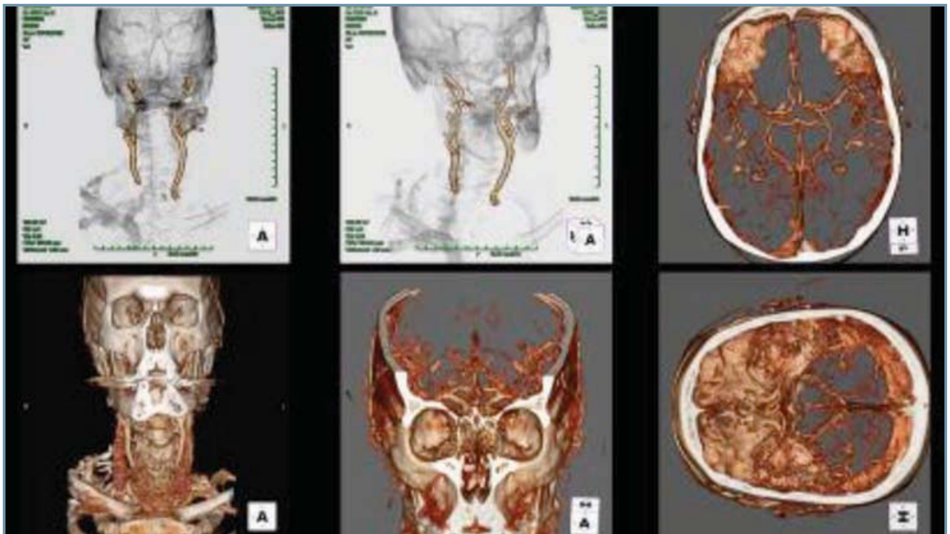


__Figure 1:__ Medical images for most scanning techniques require fine resolution and real or computer generated colour shading to show adequate detail. Accuracy and repeatability of colour representation is critical to consistent diagnosis. [Picture information on medical images courtesy of Scott Wilson, Barco]

As an aside, it is interesting to note that commercial airline pilots, who also have to read displays in high ambient light, and for whom inaccurately read data could result in disaster, are often dressed in white shirts that reflect the incident sunlight coming into the aircraft windows back onto the display panel itself and exacerbate the problems of contrast and visibility. If you want to ensure there is no adverse result due to clothing reflections dress the pilot and co-pilot in black.

Back to medical displays: Before the invention of tools such as MRI scanning, clinicians had only x-rays to use as a high-resolution diagnostic tool. The images created were produced on high-resolution photographic film that was analysed, archived and retrieved as required by different clinicians. As more imaging techniques were developed – including digital images from techniques such as position emission tomography (PET), single photon emission computed tomography (SPECT), MRI, EEG, CT or CAT – so the number of image types increased exponentially and the volume required for storage of film became a problem. The universal decision was made to store all medical images electronically and PACS (Picture Archiving and Communications System) was set up around 2000.

FOR MOST MEDICAL DIAGNOSTIC APPLICATIONS, RESOLUTIONS OF ABOVE 100 PIXELS PER INCH ARE SUFFICIENT, SO EXISTING COMMERCIAL LCDS ARE ADEQUATE

This allows retrieval of information over the health system network from the radiography department to the individual wards, and from the operating theatres to the clinics. This in turn means that the number of displays used by health services around the world has correspondingly increased exponentially with sizes ranging from ranging from 30” 6M pixel displays, medical-specific displays, down to standard commercial computer grade displays that are distributed around the general wards.

For most medical diagnostic applications, resolutions higher than 100 pixels per inch are sufficient (though not for mammography where ~ 200 pixels per inch are required), so existing commercial LCDs are adequate and are available.

When first introduced, PACS vendors produced display equipment that was optimised to reproduce the plain film of old style X-rays. CT and MRI images were reported from dedicated workstations. It is sensible to have all radiological studies reported through PACS but the plain film setup can be inadequate for CT and MRI images. High-resolution greyscale with high brightness and good contrast is a good alternative to plain films, especially mammography; but the main usage for CT and MRI requires colour, 3D and good scrolling capability to move around the multiple images that are available. Advances in LCD technology and high-power graphics cards mean that specialist medical display vendors are now able to meet all of these requirements, but at a hefty price.

In general, you would assume that the LCD cells used for these diverse applications come off much the same production lines as the normal commercial application display cells, but this is not apparent from the cost of the final medical grade product.

A 24" office monitor similar to my desktop's costs about £400 and a 30" monitor about £1,000, but a comparable 24" 100+ pixels per inch medical-grade display monitor (4M pixels) costs ~ £5K and a 200+ pixels per inch 30" display suitable for mammography (6M pixels) costs ~ £12K.

Obviously, medical displays do not come cheaply. They are specifically tailored for the application. Unlike my 30" office desktop monitor whose backlight systems shows a wide grey band on booting up the computer, medical displays need repeatable, reliable uniformity 24/7/365 to enable accurate interpretation of greyscale. Since one greyscale level can be the difference between a cancerous growth being detected by an operator or not – I am very happy that the clinicians are required to use expensive, certified, calibrated displays for this task.

The higher price for monitors built to deliver into these niche market applications that have tough specifications is a combination of small market size (i.e. number of monitors built to a particular specification) and the cost of the high specification materials and component testing and selection to maximise the performance of the monitor, such as control films, display cell, filters and backlight components.

However, though a 30" display would make for more comfortable viewing than a 24" for the radiologist, a doubling in display cost may be somewhat of a deterrent to procurement. Ideally, all medical displays should be adequate for all analysis including mammography but the x 2.5 price tag can be difficult to justify in clinical budgets.

Diagnosis and reporting of radiographic images is always performed on high-resolution monitors designed specifically for that purpose and maintained in calibration. However, back on the ward or in the clinic those images can be recalled under PACS and shown on standard (significantly lower cost), usually uncalibrated, commercial computer displays with uncontrolled lighting conditions.

Figure 2: *Greyscale images are widely used within radiography. One level of grey between adjacent pixels can make the difference between a correct diagnosis or not.* *[Picture information on medical images courtesy of Scott Wilson, Barco]*

It is acknowledged that whilst diagnosis should only be done by the radiologist or the doctor using the medical display back in the lab, the ability to recall images at will under PACS may result in decisions being made about treatment etc on displays that were not designed or sold for that purpose!

Detection of subtle changes in greyscale of the X-ray images is essential for accurate clinical diagnosis. This readability is affected by:

- Luminance of the display
- Brightness and contrast settings
- Ambient light level.

The human visual system is non-linear so its sensitivity to contrast varies with display luminance and ambient brightness. One way of defining the behaviour is by "just noticeable differences" – the smallest perceptible change in luminance that the eye can detect.

This effect has been put to use by one company, Rothband, as a way to minimise clinician error when using non-medical displays in uncontrolled conditions. They have come up with a simple but effective method to provide minimum failsafe operability that identifies potential errors in the viewing conditions caused by wrongly calibrated features such as brightness and contrast of the monitor in ambient lighting (see the figures).

At the point of logging onto the PACS system, the operator is asked to type in the random letters seen in four grayscale squares that appear in the centre of the screen. If the operator makes a mistake and types in incorrect letters, this indicates that the display contrast on that monitor at that time is insufficient for accurate viewing, and access to the images is denied. This simple process can obviate the need for expensive, time-consuming calibration of very many display devices now appearing in any health organisation.

So, is a 30" desktop monitor an indulgence for non-critical applications?

If most of your working day is spent in front of the display then it certainly is not. No matter what the task, everyone deserves to have proper equipment to perform a task comfortably. I must have a word with my accountant about upgrading my 24" display…

Chris Williams is Network Director at UK Displays & Lighting KTN (Knowledge Transfer Network)

RESISTORLESS CURRENT-MODE FIRST-ORDER ALLPASS FILTER USING CDTA

IT IS WELL KNOWN that the first-order allpass filter is widely used in several analogue signal-processing applications. In general, it is used for phase shifting from 0° to 180° (or from 180° to 0°), while keeping the amplitude of the signal constant over the frequency range of interest. It can also be used to realize universal biquadratic filters, to synthesize quadrature and multiphase oscillators, and to implement high quality factor frequency-selective filters.

On the other hand, current-mode circuits are receiving much attention because of their potential advantages such as wider bandwidth, wider dynamic range, simpler circuitry and lower power consumption. As a result, a number of current-mode first-order allpass filter realizations using different active building blocks were reported in the literature. Most of these circuits use a large passive component count and suffer from the need of passive component ratio-matching conditions. Moreover, none of them are electronically adjustable. Although first-order translinear-C current-mode allpass sections with electronic tuning properties were reported in S Maheshwari and I A Khan's *"Simple first-order translinear-C current-mode allpass sections"* in Int. J. Electron., Vol 90, pp79-85, 2003, they suffer from low output impedances.

In this article, an electronically tunable current-mode first-order allpass filter realization using only two CDTAs and a single-virtually grounded capacitor is introduced. The proposed circuit can realize both inverting and non-inverting types of the current-mode first-order allpass filter, without changing the circuit topology.

Due to electronically tenability properties of the CDTA, the phase response of the proposed circuit can be adjusted by an external bias current. No component-matching condition for realizing the allpass function is required. The circuit also exhibits high-output impedance, which is easy cascading in the current-mode operation. Simulation results verifying theoretical analyses are included.

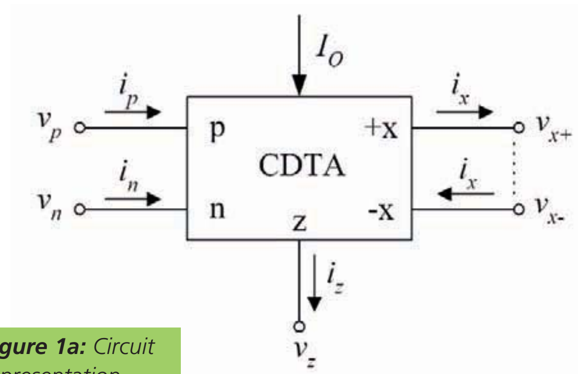


Figure 1a: *Circuit representation*

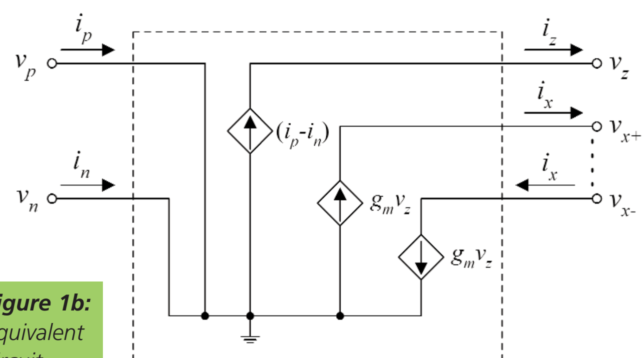


Figure 1b: *Equivalent circuit*

Circuit Description

As shown in **Figure 1**, the CDTA is a versatile current-mode active building block, which has the following port relations as follows:

$$v_p = v_n = 0,\ i_z = i_p - i_n \text{ and } i_x = g_m v_z \tag{1}$$

In this equation, g_m is the transconductance gain of the CDTA, which can be defined by:

$$g_m = \frac{I_O}{2V_T} \tag{2}$$

where $V_T \cong 26\text{mV}$ at 27°C is the thermal voltage and I_O is the external bias current of the CDTA.

Figure 2 shows the proposed resistorless current-mode first-order allpass filter with electronic tuning properties, employing only two CDTAs and one virtually grounded capacitor. A routine analysis of the proposed circuit yields the following current transfer functions:

$$\frac{I_{AP+}}{I_{in}} = -\frac{I_{AP-}}{I_{in}} = \frac{1 - s\left(\dfrac{C}{g_{m1}}\right)}{1 + s\left(\dfrac{C}{g_{m1}}\right)} \tag{3}$$

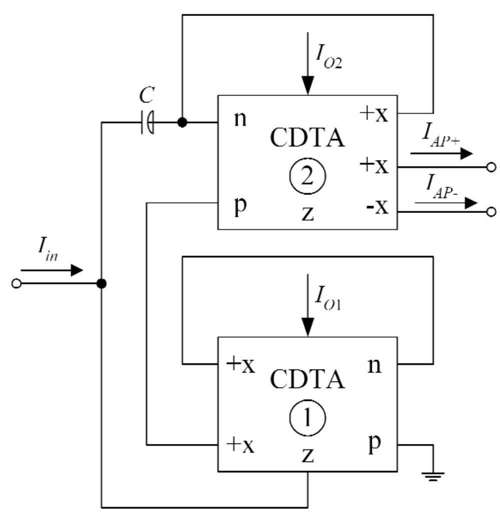


Figure 2: *The proposed CDTA-based, current mode, first order, allpass filter*

From **Equation 3** it can be seen that the circuit can realize both inverting and non-inverting type first-order allpass functions without changing the circuit configuration. As it is demonstrated in Figure 2, the CDTA-based circuit does not require any external passive resistor and does not require any matching conditions for realizing first-order allpass function. Here, the pole frequency of the circuit is expressed as:

$$\omega_o = \frac{g_{m1}}{C} = \frac{I_{O1}}{2V_T C} \quad (4)$$

and the phase responses are respectively given by:

$$\phi_{AP+} = -2\tan^{-1}\left(\frac{\omega C}{g_{m1}}\right) \quad (5)$$

and $$\phi_{AP-} = 180^\circ - 2\tan^{-1}\left(\frac{\omega C}{g_{m1}}\right) \quad (6)$$

Equations 5 and **6** show that the proposed filter can provides phase shifting both between 0° to -180° and 180° to 0°. Moreover, the shifted phase value can be controlled electronically by adjusting I_{o1}.

Simulation Results

A PSPICE program was carried out to check the workability of the proposed circuit. In simulations, the CDTA was constructed as described in *"Multiple-input single-output current-mode multifunction filter using current differencing transconductance amplifiers"* in Int. J. Electron. Vol 61, pp209-214, 2007 by W Tangsrirat, T Dumawipata and W Surakampontorn, with the transistor model parameters of PR100N (PNP) and NP100N (NPN) and DC supply voltages of ±3V. The designed capacitor value and the bias currents of the CDTAs were chosen as C = 1nF and $I_{o1} = I_{o2}$ = 100μA.

Figure 3 shows the time domain responses of the proposed filter in which 100μA peak sinusoidal input current at 318kHz is applied. The simulated phase-frequency plots of both types of the allpass filter for three different bias current values are depicted in **Figure 4**. It is observed from the figure that the electronically tunability of the phase response can be achieved through I_{o1}.

Conclusion

In this study, a canonical current-mode first-order allpass filter is presented. It can realize both inverting and non-inverting type allpass filtering functions by using only two CDTAs and one virtually grounded capacitor.

The proposed circuit has following advantages: (a) the phase shift can be electronically tuned by an external bias current; (b) no matching condition is imposed for realizing the allpass functions; and (c) since all the outputs are provided through high output impedance terminals, it can be directly connected to the next stage without any impedance matching requirement.

Worapong Tangsrirat
Thailand

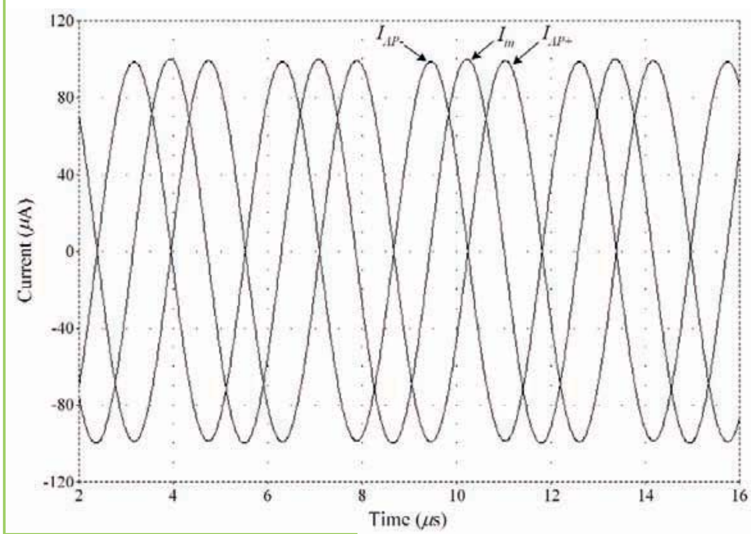


Figure 3: *Time domain responses of the proposed allpass filters*

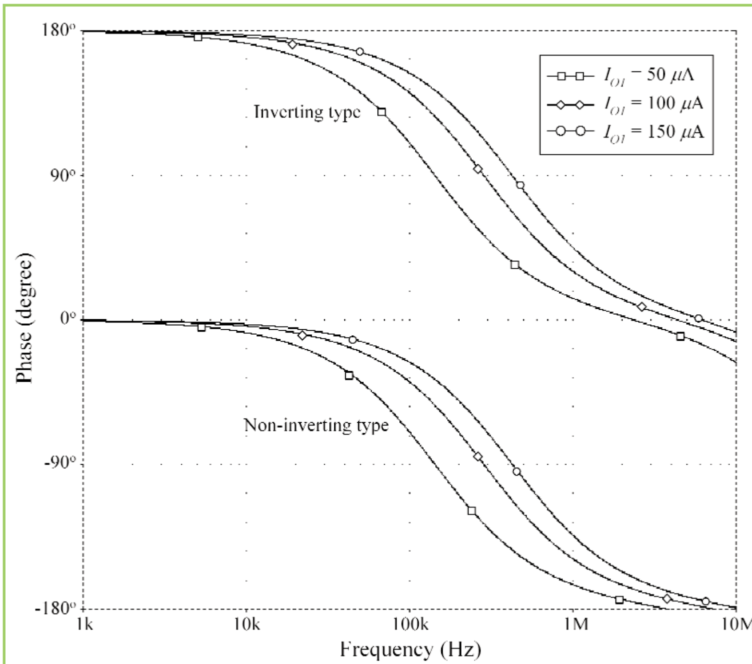


Figure 4: *Simulated phase responses of the proposed allpass filters when I_{o1} is varied.*

TIP 1: TRIPLE OUTPUT LED DRIVER WORKS WITH COMMON ANODE LED STRINGS

By Hua (Walker) Bai, Linear Technology

Some multi-string LED modules come with a common anode figuration, whereas the common anode connection reduces the number of wires between the LED module and its driver from 2N to N+1, where N is the number of LED strings in the module. This idea illustrates how to drive such a common anode LED module while simultaneously limiting the LED string voltage when an LED string becomes open.

Figure 1 shows the LT3496 triple output LED driver in a buck mode configuration, where the LED strings reside between PVIN and the 200mΩ sense resistors to allow the common anode connection at PVIN. This is in contrast to the usual buck mode configuration for three free-floating LED strings. In a normal steady state operation, of current this circuit delivers 500mA to each LED string.

The programmed overvoltage protection (OVP) is not always needed in a buck mode LED driver circuit. Unlike boost, buck-boost and SEPIC drivers, the switch voltage of a buck mode LED driver droops when an LED string is opened. In this case, OVP is not needed. However, the CAP1 pin can be used as an open circuit indicator. Furthermore, an open collector buffer may be needed in some applications. For simplicity, the reference designators of Channel 1 are used exclusively (in the description below).

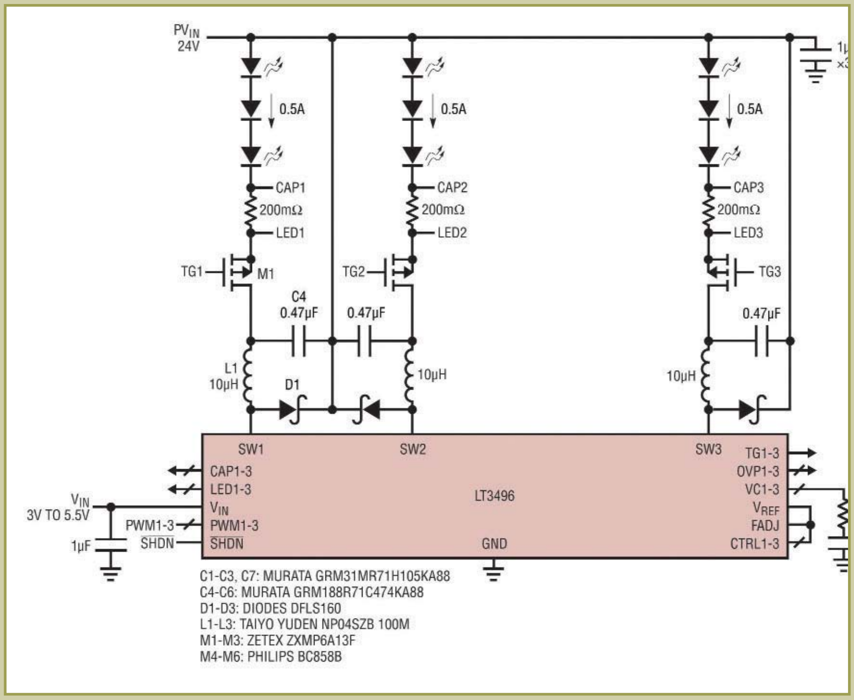


Figure 1: *A triple buck mode LED driver works with common anode LED strings*

A potential issue arises if an LED string goes open and is subsequently reconnected. For instance, this could occur if a cable connection between the LED driver and the LED module is not a constant connection and intermittently disconnects and reconnects. Under these circumstances, the LED string can experience a large inrush current for a number of microseconds after it is reconnected. This large current is due to the discharge of capacitor C4. The amplitude of this inrush current is related to the difference between PVIN and the LED string voltage – the larger the difference, the higher the inrush current. In Figure 1, for example, a 24V input and the 3-LED configuration, shows the measured inrush current peaks at 1.2A.

If inrush current is a concern, then the voltage across the LED string terminals needs to be clamped to a voltage that is just slightly higher then the LED string voltage when the string is open.

Figure 2 shows a circuit that limits the voltage across the LED string to an OVP level set by resistors R1 and R3. This would be 15V in this example. However, in order for an OVP circuit to be effective, CAP1 must be brought up after the OVP logic turns off the main switch. Resistor R4 provides a few hundred micro-Amps of pull up current for CAP1. Without R4, CAP1 is held low, thus making the OVP circuit moot.

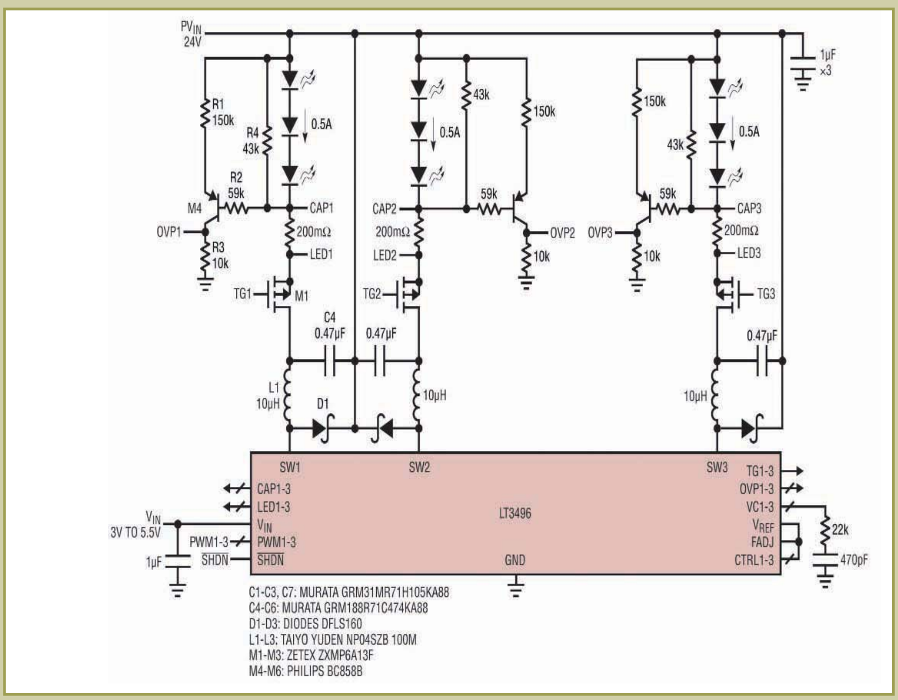


Figure 2: *A triple buck mode LED driver has open LED protection*

OMNETICS'S OVER-MOLDING CAPABILITIES EXTENDED TO NEW LATCHING NANO CONNECTOR

Omnetics Connector Corp, manufacturer of miniature high rel connectors, has announced that its unique, latching Bi-lobe connector can now be offered with overmolded cables, or incorporated into equipment housings. This offers several benefits for medical applications, including protection against contamination and easy sterilization, as well as significant savings in space, weight and cost.

Omnetics's 0.025in pitch latching Bi-lobe connector is unique in that the mechanism provides very high levels of mating security – devices pass the shock and vibration requirements of MIL-DTL-32139 – yet connector halves simply push together requiring no tools or complicated procedures.

Unlike other suppliers who require large production orders to undertake customized over-molding, Omnetics can offer a precision over-molding service for the latching Bi-lobe and many other miniature connector families even for small order quantities. A wide range of materials, from soft silicones to hard glossy shells, can be used.

Omnetics was formed in 1984 to deliver rugged, reliable interconnect solutions for the most demanding industries. The company has a fully integrated design and manufacturing plant in the US, where it produces micro and nano miniature interconnect products, featuring COTS, standards and custom connectors for various industries.

www.omnetics.com

HIGH EFFICIENCY, COMPACT, LED DRIVER IC

The new A6210 from Allegro MicroSystems Europe is an LED driver IC based on a buck (stepdown) regulator circuit using constant 'on' time with valley current-mode control. This control scheme allows the use of very short switch 'on' times, making the new device ideal for applications requiring high switching frequencies combined with high input voltages and low output voltages.

The small size and high efficiency of the A6210 makes it suitable for applications where space is limited, while its high current and medium-voltage capability mean that it can also be used in high-power applications such as street lighting.

System cost is reduced as a result of the high switching frequencies (up to 2MHz), which allow small, low-value ancillary components, such as inductors and capacitors, to be used. In addition, fewer external components are required because of the high levels of integration in the device, which includes a 3-amp switch. Optimal drive circuits are used to minimise switching losses, allowing over 90% efficiency in realistic conditions.

The switching frequency is maintained constant, as the input voltage modulates the 'on' time, which ensures excellent line correction.

www.allegromicro.com

DC/DC CONVERTERS WITH REINFORCED INSULATION FROM POWERSOLVE

Powersolve announces the THI-2M series of high isolation DC/DC converters designed to provide 2W in a small DIP-16 package. Input/Output-isolation voltage is specified for 4000 VACrms, which qualifies the converters for applications that request a supplementary or reinforced insulation approved to EN/UL 60950-1 industrial or EN/UL 60601-1 medical safety standard.

The THI-2M series features models with input voltages of 5, 12 or 24VDC and regulated output voltages of 5, 12, 24, ±12 and ±15VDC. Operating temperature range is specified for -25°C to +71°C. The product is also available in a SMD package version (TES-2M series).

The THI-2M converters provide an ideal solution for many cost-critical applications in instrumentation or in commercial and medical equipment.

Powersolve Electronics Ltd was established in 1987 as a distributor of power products and as a manufacturer of standard and custom power solutions.

All products supplied by Powersolve are manufactured in ISO 9000 approved facilities and carry Full International Safety Approvals.

www.powersolve.co.uk

DIGITAL MULTIMETERS WITH COMPREHENSIVE FUNCTIONS FOR ELECTRONIC AND INDUSTRIAL USE

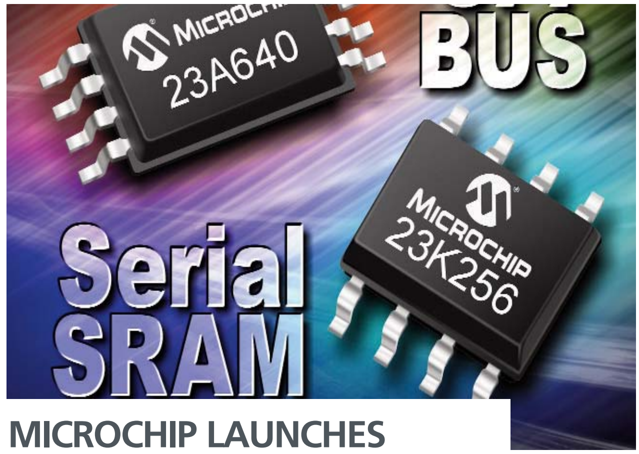
The new Yokogawa TY700 Series is a range of handheld digital multimeters that combine high accuracy, performance and reliability with a range of new functions for both electronics and general industrial applications.

Featuring a basic accuracy of 0.020% and a 50,000 count dual display with a 51-segment bargraph, the new instruments offer true RMS measurement and simultaneous measurement and display of DC and AC as standard, as well as full support for data management via a large-capacity logging-mode memory and USB communication.

There are two models in the TY700 Series: the Standard Type TY710 with 20kHz bandwidth and the Advanced Type TY720 with 100kHz bandwidth. Additional features on the TY720 include switchable RMS/mean detection, a low-pass filter, a 50mV AC range and low-power resistance measurement. The TY720 also has a logging-mode memory with a capacity of 10,000 data points, compared with 1000 for the TY710. Both products are offered with USB communication and an optional communications software package.

The instruments conform to the EN61010-1 safety standard (1000 V CAT III, 600 V CAT IV) and have safety shutters for preventing the erroneous insertion of the test leads into the current measurement terminals.

www.yokogawa.com



MICROCHIP LAUNCHES STANDALONE SERIAL SRAM DEVICES

Microchip announces a family of 8kByte and 32kByte standalone serial SRAM devices designed to increase a system's available RAM through adding small, inexpensive external devices. The 23A640, 23K640 (23 x 640), 23A256 and 23K256 (23 x 256) devices feature a familiar, industry standard SPI interface, providing increased design flexibility while reducing design and production costs.

Many embedded applications require volatile RAM for temporary data storage, or for use as a scratchpad, for bulk processing and for math algorithms. In many cases, this RAM is embedded within the microcontroller (MCU). In the past, the most viable way to add more RAM was to buy a larger MCU, which could add unnecessary feature overhead and increase design costs. The only alternative was to add large, parallel-access RAM devices that use up large numbers of I/O pins.

Microchip's serial SRAM devices provide a simple, inexpensive way for designers to add more RAM to their application while keeping the same MCU or, as they require fewer MCU I/O resources, even using a smaller MCU. The serial RAM devices require just four I/O pins as opposed to 16 or 24 pins for a parallel RAM.

www.microchip.com

SCHROFF PRODUCTS CONFORM TO NEW IEC STANDARD FOR 1U CHASSIS

Electronics packaging specialist Schroff has announced that the 1U versions of its multipacPRO chassis and CompactPCI and VME64x systems all conform to the new IEC 60297-3-105 standard for 1U-high 19in chassis.

Covering 'Dimensions and design aspects for 1U chassis', the IEC 60297-3-105 standard defines three types of 1U-high chassis: Type A are connected to the 19in cabinet or frame at the 19in front panel only, Type B are connected to the 19in cabinet or frame at the 19in front panel and are further reinforced by means of slide rails, and Type C are connected to the 19in cabinet or frame via telescopic rails.

Both the steel and aluminium versions of Schroff's 1U-high multipacPRO chassis comply with IEC 60297-3-105 and can accommodate non-standard components and power supply units or standard circuit board formats using an internal mounting kit.

Also fully compliant are Schroff's 1U-high 19in CompactPCI and VME64x systems, which are based on the company's europacPRO subrack. Equipped with appropriate backplanes, ventilation units and IEEE guide rails, these shielded systems can accommodate two horizontal 160mm-deep boards, as well as horizontal rear I/O boards.

www.schroff.co.uk

THE AXIOHM SENS' N PRINT ECO FRIENDLY TICKET PRESENTER FROM DED LIMITED

Now available from DED Limited, the Axiohm Sens' n Print delivers an eco friendly ticket solution with all the advantages of a ticket presenter/retractor, but with substantial cost savings. Sens' n Print promises to reduce costs and improve the reliability of kiosks by removing the need for a paper presenter without compromise on quality or security.

With the opportunity to minimise paper use and be a friend to the environment, Sens' n Print works in a very unique way. Once a print job is sent to the printer, Sens' n Print holds it and waits until movement is detected near the receipt outlet before printing. If no movement is detected after a pre-determined amount of time, the print job is cancelled, no receipt is printed and so paper consumption is reduced.

The Sens' n Print unit is ultra compact and when combined with no requirement for ticket presenters, retractors or paper bins, offers the advantage of dramatically reducing the space required inside a kiosk. A combination of Axiohm's printer controller board, a high tech move sense antenna and minimal components makes Sens' providing the opportunity to simplify kiosk design.

www.ded.co.uk

DIN-RAIL POWER SUPPLIES IDEAL FOR "DIFFICULT" INDUSTRIAL ENVIRONMENTS

uPowersolve announces a new family of 90 to 600W DIN-rail power supplies, designed for use in a wide variety of applications including "difficult" industrial environments. The Tracopower TSP Series units feature a rugged metal case construction designed to be shock and vibration proof and offer excellent electrical specifications and immunity against electrical disturbances. Installation is made easy thanks to detachable screw terminal blocks and snap-on DIN-rail mounting.

Twelve standard models are available with tightly regulated, adjustable output voltages of 12, 24 and 48VDC, with ratings up to 25A. The DIN-rail power supplies are specified for industrial operating temperatures from –25°C to +70°C with convection cooling and also provide thermal overload protection. Remote on/off is standard.

The TSP Series also includes three optional add-on Function Modules that offer battery back-up, output buffering or true redundant operation in system applications.

The TSP Series has worldwide safety approvals to EN60950-1, UL508 and EN60079-15 standards and EMC Compliance to EN61204-3 standards for industrial power applications.

These new power supplies are ideal for powering sensitive loads in industrial process control systems, machine tools and other demanding industrial applications.

www.powersolve.co.uk

KEEPING THINGS SIMPLE – ONE FOR ALL

The HMP series includes three programmable high-performance power supplies, featuring two or three output channels that cover ranges of 0 to 32 volts with up to 10 amps. New functionalities, a high set and read-back resolution, the EasyArb feature and excellent noise/ripple values (150µV) characterize the family of power supplies.

Hameg has set great value on the logically combinable electronic fuses (FuseLink), as well as on a convenient tracking function.

The high read-back resolution, down to 1mV/0.1mA, meets even the most stringent requirements. Moreover, applying the EasyArb function, users can form arbitrary voltage/current shapes for each channel.

The HMP series is provided with a LCD display, a dual serial interface (USB/RS-232) (optionally Ethernet/USB or GPIB) and additional terminals for all channels on the rear panel.

Available from February 2009, the HMP series will go for a list price starting from EUR 959.

www.hameg.com/HMP2020

KEITHLEY RELEASES NEW 2009 TEST AND MEASUREMENT PRODUCT GUIDE

Keithley Instruments announces the release of its 2009 Test and Measurement Product Guide. The 144-page guide offers details and technical specifications on Keithley's general-purpose and sensitive sourcing and measurement products, DC switching, RF switching and measurement, data acquisition solutions, semiconductor test systems and optoelectronics test. A useful selector guide simplifies choosing the right solutions for specific applications. For a free copy of Keithley's 2009 Test and Measurement Product Guide, visit http://www.keithley.info/catalog09.

Keithley's 2009 Test and Measurement Product Guide is arranged by product type and application area with sections containing the newest innovations in test and measurement, including: digital multimeters (DMMs) and systems, switching and control, RF/microwave and wireless switching and testing and measurement, specialized power supplies, source and measure products, optoelectronic test, function/pulse/arbitrary function generators, semiconductor test, low-level measurement and sourcing and others.

www.keithley.com

PRO-WAVE'S CAR-REVERSING SENSORS NOW AVAILABLE FROM LPRS

LPRS, manufacturer of easy-Radio, now offers a wide range of Pro-Wave components for proximity detecting applications.

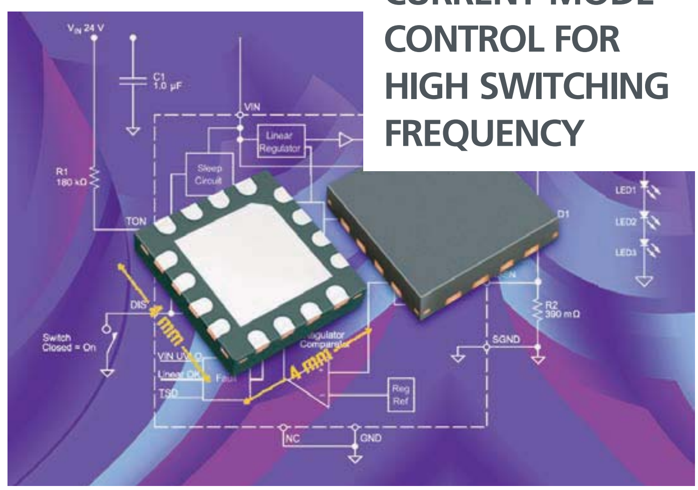
The Pro-Wave SRM400 sonar ranging modules and US040 sensors are widely used for car reversing aids and are very popular with volume manufacturers and after market suppliers.

The SRM400 is a sonar ranging module utilising the newly developed Sonar Ranging IC, PW-0268, which can work with all of Pro-Wave's PT or EP type transducers. The SRM400 provides a time-saving shortcut for design engineers who may have limited experience of analogue design and/or the operation of ultrasonic transducers. When developing car reversing systems or other distance measurement products use of this ready-to-go module allows engineers to focus their efforts on the system integration, digital processing, software and mechanical design.

The Pro-Wave US040015 is an ultrasonic sensor generates asymmetrical beam patterns, approximately ±80° about the horizontal axis and 40° about the vertical, which meets the requirements of most car reversing aids. Thanks to the newly developed Sonar Ranging Chip PW-0268 the US040015 sensor is ultra-compact aiding installation in difficult locations.

www.lprs.co.uk

BUCK REGULATOR IC USES VALLEY CURRENT-MODE CONTROL FOR HIGH SWITCHING FREQUENCY



The new A4403 from Allegro MicroSystems Europe is a 3A buck regulator IC which uses a valley current-mode control scheme to offer very low switch-on times, making it ideal for applications that require high switching frequencies combined with high input voltages and low output voltages.

The high step-down ratio made possible by an 'on' time of less than 50ns permits a wide output voltage range from 0.8V to 15V at 3A from an input supply voltage range from 9V to 46V. Standby current is less than 100 microampere.

The device's capability for high switching frequencies (up to 2MHz) leads to the use of smaller and lower-value ancillary components such as inductors and capacitors, while the high level of integration means that a minimal number of external components are needed. As a result, the A4403 provides a compact, flexible and cost-effective solution for office automation, industrial and consumer applications.

Optimal drive circuits are utilized to minimise switching losses and the switching frequency is maintained constant as the input voltage modulates the 'on' time. This feed-forward control ensures excellent line correction.

www.allegromicro.com

MIKM MINI CIRCULAR CONNECTORS OFFER HIGH PERFORMANCE UNDER EXTREME CONDITIONS

A new series of mini circular connectors developed by leading hi-rel connector company, ITT Interconnect Solutions, provides a reliable interconnect solution capable of surviving the harsh environment typically found in aircraft. MIKM mini circular connectors can meet challenging conditions including extreme temperature fluctuations and provide highly reliable signal fidelity under shock and vibration loads.

The MIKM mini circular series includes 7, 55 and 85 contact positions on 1.27mm (0.050in) contact spacing and feature ITT's twist pin contact system. Secure mating is assured via a threaded coupling mechanism fully terminated to crimped gold-plated contacts. Size 24 to 32 AWG wire sizes are available. Extremely robust yet cost-effective, the connectors have thermoplastic dielectric insulation and are available in reverse gender configurations.

Originally developed by the company for its customers in the aircraft industry, the MIKM mini circular series suits a wide range of applications, including medical electronics, tactical communications and radar systems..

www.ittcannon.com

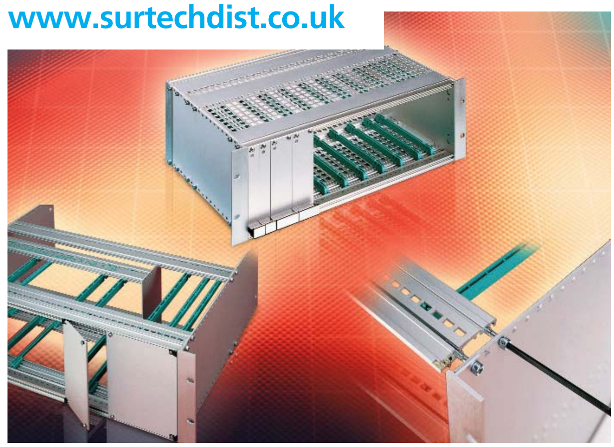
MARKET-LEADING KM6-II SUBRACKS STILL AVAILABLE FROM SURTECH

Surtech Distribution, the enclosures and accessories specialist, has announced that it has been appointed main UK distributor of the "gold standard" in subracks: the KM6-II, manufactured by HDD of Bremen, Germany. The KM6-II is well-established as the market leader, renowned for having an unbeatable combination of versatility and cost-effectiveness: more the subrack has been continuously updated in line with technological advances.

Surtech can supply the KM6-II in kit form or fully assembled, configured to specific customer requirements. Technical help is available through Surtech's dedicated technical sales team, who have the expertise to advise customers on the most cost-effective product for their particular needs. Additionally, Surtech offers a complete modification and customization service for KM6-II subrack front panels to suit specialised or unusual applications.

KM6-II subracks are available in two versions: KM6-II Standard, which is optimized for 160mm and 200mm, 3U and 6U eurocards; and KM6-II Universal, which is suitable for 3U, 4U, 6U and 9U applications. With end plates up to 420mm high, the KM6-II Universal is extremely versatile and can be configured in many different ways as there are hundreds of parts and accessories available. It is also unquestionably the strongest subrack on the market, with two screw fixings on all tie-bars, enabling it to withstand the weight of the heaviest components, as well as shock and vibration.

www.surtechdist.co.uk



NEW COM EXPRESS MODULE WITH AN INTEL ATOM PROCESSOR FROM CONGATEC

Congatec AG has launched its conga-BA945, the fourth embedded computer module based on the Intel Atom processor family. This is a COM Express module with type 2 pinout and conforms to the specified COM Express basic size of 95 x 125mm².

The Intel Atom processor N270 has a clock speed of 1.6GHz, as well as 512kB cache and a 533MHz frontside bus speed. Despite the high processing power, the processor gets by with a maximum power dissipation of 2.5W. Thanks to the sophisticated power management provided by the congatec embedded BIOS, the actual values reached during practical use remain well below this maximum level.

The congatec conga-BA945 supports Intel Hyper Threading Technology, which means with the appropriate software it has the capability of running two operating systems in parallel and completely independent of one another.

In contrast to the recently announced conga-CA945, which is an even lower priced and more compact COM Express compatible module featuring an Intel Atom processor, the conga-BA945 makes use of the more powerful Intel 945GME chipset versus the Intel 945GSE used on the conga-CA945.

www.congatec.com

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54600B	Dual Trace 100MHZ 20 MS/S	£225

PHILLIPS

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PM3065	2+1 Ch. 100MHZ dual TB/Delay. Autoset	£125
PM3084	4 Ch. 100MHZ Delay TB	£140
PM3335	Digital Storage Dual Trace 60MHZ 20 MS/S	£125
PM3337	Rack Mount Digital Storage Dual Trace 60MHZ 20 MS/S	£125
PM3365A	Digital Storage Dual Trace 100MHZ 100 MS/S	£150

HITACHI

V222	Dual Trace 20MHZ DC Offset, Alt. Mag	£50
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Increased Focus on Power Supplies and DC/DC Converters

Arrow Electronics has expanded its power technology portfolio by signing a distribution agreement with Recom, a leading global supplier of DC/DC and AC/DC converters. Arrow will sell and support Recom's complete power converter product range throughout Europe.

The new agreement is in line with Arrow's strategy to provide the widest possible range of power supply and converter technologies to designers across Europe. For Recom, the agreement will expand the penetration of its product ranges into European markets and ensure that customers have access to products and support at a local level.

Available in single, dual, and triple rail outputs and a wide variety of input and output voltage configurations, power ratings, and isolation voltages, Recom converters are used in most industries requiring low- to medium-power voltage conversion in the 0.25W to 60W range. The company also offers the largest range of safety agency-approved converters from any manufacturer.

www.arrowne.com

Harwin in an Exclusive European Trio-Tek Catalogue Sales Agreement with Farnell

High-reliability interconnect company Harwin has appointed Farnell as exclusive European catalogue distributor for its innovative Datamate Trio-Tek range of crimp contact connectors.

Harwin's Datamate Trio-Tek open-barrel crimp contact offers significant advantages over conventional Datamate contacts whilst providing the same high reliability, as it enables assembly time to be shortened significantly and therefore reduces processing costs. The new crimp design features a triangular form that simplifies the insertion of contacts into the housing, enabling customers to fully automate crimping in medium and high volume applications.

Harwin's Datamate Product Manager Graham Cunningham commented: "Farnell is one of Europe's leading distributors and justifiably known for offering customers an extensive range of the best products and latest technology from leading manufacturers. We have a strong relationship with them and believe their reputation for first-class customer service and the strength of their catalogue media will provide an excellent partner for Trio-Tek. We are delighted to have appointed them as our catalogue distributors for this important initial phase of Trio-Tek sales in Europe."

Farnell's Justin Willoughby said: "We have a long and fruitful relationship with Harwin and always welcome the opportunity to offer their latest products to our customers. Harwin's expertise as a global leader in the design and manufacture of electronic interconnect solutions means their products are always leading edge and provide elegant solutions for electronics engineers across a great number of applications and markets."

www.farnell.co.uk
www.harwin.co.uk