A.Y. Nikolaenko, A.I. Strokov, P.A. L’vov*, E.A. Moiseykina*

RFID BASED SORTING AND STACKING SYSTEM FOR ASSEMBLY LINES

(Yuri Gagarin State Technical University of Saratov,
OJSC “Signal” Engels Design Bureau named after A.I. Glukharev)

New technologies have a great influence on the production process in modern factories. Introducing new techniques and methods is crucial to optimize and enhance the working of factories. However, ensuring a reliable and correct integration requires complete evaluation and assessment. Developers use RFID systems to develop and implement real time solutions to enhance and optimize the production and assembly processes.

In the paper [1] the authors built an RFID based stacking-transport system. The system was applied on a production line shown in Figure 1. A high frequency RFID tag was attached to each box. It contained information about its product brand. A high frequency reader reads tags on coming boxes and sends the product brand information to a sorting controller which in turn determines where to route the box. A stacker crane robot stacks forwarded boxes on pallets according to their brand. Pallets also have RFID tags. At the end of the assembly line another reader that uses another frequency updates the tag of each pallet, which is now full with boxes, with the new information about the contained boxes. The paper compares the system to the existing barcode based systems. The former enhances time management and reduces the number of damaged products because it needs no manual scanning of tags. Also, the memory property allows better management of the products in the warehouse.

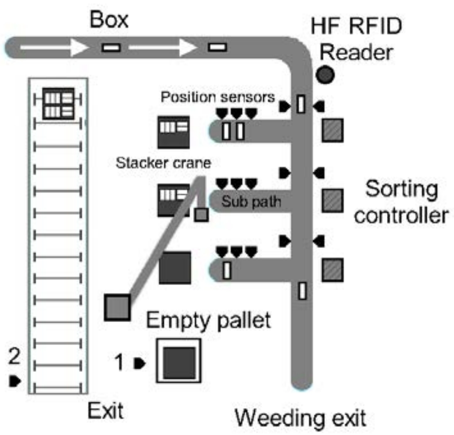

Figure 1. Scene in production line

The read/write range performance of a RFID system depends mainly on the choice of frequency, radiated power from the reader, sensitivity and modulation efficiency of a transponder, data rate, reader receiver sensitivity in the presence of self-jammer signal and location of the transponder [2]. The biggest challenge for the receiver front-end is to handle leakage from the full power continuous wave (CW) signal being transmitted during reception to keep the passive transponders powered up. This calls for the design of a wide dynamic range receiver or a use of some sort of an isolation approach, or self-jammer cancellation technique. However, the isolation between transmitting and receiving channels increases the RFID reader cost. Leakage canceller complicates the reader receiver and adds to the reader’s consumption that is critical for mobile applications [3], [4].

This paper proposes a new RFID reader based on automatic network analyzers (ANA) that conventionally has been applied for microwave load’s complex reflection coefficient measurement [8], [9]. This approach does not require the signal reader carrier’s compensation. ANA based on down conversion of measuring frequency of multi-port reflectometer ports. The proposed RFID reader block-diagram is shown in Figure 2. The technique consists in use of MR in combination with down conversion of measurement frequency. The outputs of measuring ports of MR are connected to the mixers of frequency conversion unit instead of power meters. After the heterodyning the analogue audio frequency signals proportional to probes responses are sampled by the data acquisition board (DAB) and entered to computer memory. All the subsequent processing of the obtained data is performed in the digital form using the special software.

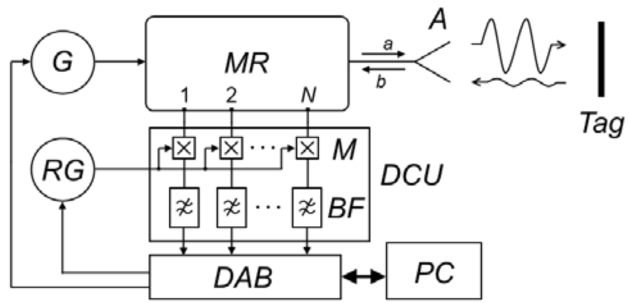

Figure 2 The block-diagram of proposed ANA based RFID reader: *G*, *RG* are microwave generator and reference generator; *MR* is multi-port reflectometer; 1, 2, ..., *N* are measuring ports; *A* are antenna; *M* are mixers; *BF* are band-pass filters; *DCU* is down conversion unit; *DAB* is data acquisition board; *PC* is personal computer.

Here the output digital signals of the DAB are linear functions of the estimated parameters *a* and *b* so they may be processed without the loss in signal-to-noise ratio and parameters of the load under test can be measured with potentially achievable accuracy.

Let the output signal of the j-th probe of MR to be applied to the first input of corresponding mixer, the sine voltage of the reference generator RG being impressed at the second. Then the output voltage of the j-th mixer can be represented as a function of time:

$$\begin{aligned} \upsilon_j &= U_j \cos(\omega \cdot t + \varphi_j) \cdot \cos[(\omega - \nu)t] = \\ &0{,}5 \cdot U_j \{\cos(\nu \cdot t + \varphi_j) + \cos[(2\omega - \nu)t + \varphi_j]\}, \quad j=1,2,...,N \end{aligned} \qquad (1)$$

where U_j, ϕ_j are unknown amplitude and phase of the microwave signal in the j-th measuring port of MR respectively; ω is the known frequency of the microwave signal; $\Delta\omega$ is the difference between the frequencies of the main G and reference RG generators; N is the number of measuring ports of MR.

The band-pass filter extracts only low frequency component of the signal $\upsilon_j(t)$:

$$\vartheta_j(t) = U_j \cos(\nu \cdot t + \varphi_j) + \xi_j, \quad j=1,2,...,N \qquad (2)$$

where ξ_j is the measurement error at the j-th channel. After the analogue-to-digital converter of the DAB the signals v_j are digitized forming the N series each of K samples:

$$\vartheta_j(t_k) = \vartheta_{jk} = U_j \cdot \cos(\nu \cdot k \cdot \tau + \varphi_j) + \xi_{jk}, \quad (j=1,2,...,N;\ k=1,2,...,K) \qquad (3)$$

where τ is the sampling period (the time between two sequential samples in one series); k is the sample number in the series (in each of N measuring channels the series of K digital samples is obtained).

The problem is to determine the estimates of unknown parameters a and b using the digital voltage measurements v_{jk} (3). It is well-known [5] that the complex amplitude of the j-th MR port response u_j is related to complex amplitudes of incident b and reflected a waves with the equation

$$u_j = A_j \cdot a + B_j \cdot b + \Xi_j, \quad j=1,2,...,N \qquad (4)$$

where A_j, B_j are known complex gains of the j-th channel for incident and reflected waves respectively found after the calibration procedure; δ_j is the error of measurement of u_j. The responses of MTLR have the same form (4), but the magnitudes of coefficients $|A_j|=|B_j|$ and phase differences $\psi_j=arg(B_j/A_j)$ are implied to be exactly known. Thus, if the estimates of complex amplitudes u_j are known one can calculate the estimates of parameters a and b characterizing the termination under test. Therefore, the first stage of processing of the data v_{jk} obtained from the measuring ports should be aimed at estimation of complex amplitudes u_j. After the variable substitution

$$\begin{cases} y_j = U_j \cos(\varphi_j), \\ z_j = U_j \sin(\varphi_j), \end{cases} \quad \begin{cases} x_{1k} = \cos(\nu \cdot k \cdot \tau), \\ x_{2k} = -\sin(\nu \cdot k \cdot \tau), \end{cases} \quad (j=1,2,...,N;\ k=1,2,...,K) \qquad (5)$$

the set (3) can be rewritten in the form:

$$v_{jk} = y_j \cdot x_{1k} + z_j \cdot x_{2k} + \xi_{jk}, \qquad (6)$$

It is quite natural to assume that the frequency difference $\Delta\omega$ remains constant during the measurement period K_τ. That is why all the parameters $\Delta\omega$, x_{1k} and x_{2k} implied to be known in every measurement. And the errors ξ_{jk} occur largely owing to the thermal noise of matching amplifiers of the DAB hence they may be considered as the independent sample of the normal random process with zero mathematical expectation and unknown fixed variance σ^2. It is evident from (5) that unknown values of y_j

and z_j are real and imaginary components respectively of corresponding complex amplitudes u_j. The estimates of these components can be determined from the linear set (6) using the maximum likelihood (ML) method. If the assumptions made about errors ξ_{jk} are true the named method consists in solution of (6) by the least square (LS) procedure. Thus, the estimates of y_j and z_j are calculated from the expression

$$\hat{\boldsymbol{u}}_j = (\boldsymbol{X}^T \boldsymbol{X})^{-1}(\boldsymbol{X}^T \boldsymbol{V}_j), \quad j=1,2,...,N \qquad (7)$$

where $\hat{\boldsymbol{u}}_j = (\hat{y}_j, \hat{z}_j)^T$ is the vector of corresponding estimates; $\boldsymbol{V}_j = (v_{j1},...,v_{jK})^T$ is the vector consisting of K samples of voltage in the j-th channel; ^{T} and $^{-1}$ designate the operations of matrix transpose and inversion respectively; X is the matrix of design of experiment

$$\mathbf{X}^T = \begin{bmatrix} x_{11} & x_{12} & \dots & x_{1K} \\ x_{21} & x_{22} & \dots & x_{2K} \end{bmatrix}$$

After all the estimates $\hat{y}_j$ и $\hat{z}_j$, are found it is easy to form the estimates of complex amplitudes $\dot{u}_j$

$$\hat{u}_j = U_j \cdot \exp\{i \cdot \varphi_j\} = \hat{y}_j + i \cdot \hat{z}_j, \qquad (8)$$

where i is the unit vector determining the direction of imaginary axis on a complex plane (unit imaginary number). Since the errors ξ_{jk} are normally distributed and independent then the real and imaginary parts of errors δ_j from the set (4) will be distributed normally as well. Substituting estimates of $\dot{u}_j$ from (8) to (4) one can get the set of equations to be solved for unknown variables a and b which is similar to the set (6). The only difference between (4) and (6) is that all the parameters in the first set are complex. It is known [6] that the optimal estimates of unknown parameters can be obtained from the solution of (4) by the ML method consisting in calculation the LS estimates provided the errors δ_j being normally distributed

For the sake of simplicity the set of N complex equations is transformed into equivalent set of $2N$ real ones.

$$\begin{aligned} \hat{y}_j &= \mathrm{Re}(A_j)\mathrm{Re}(a) - \mathrm{Im}(A_j)\mathrm{Im}(a) + \mathrm{Re}(B_j)\mathrm{Re}(b) - \mathrm{Im}(B_j)\mathrm{Im}(b) + \mathrm{Re}(\Xi_j), \\ \hat{z}_j &= \mathrm{Im}(A_j)\mathrm{Re}(a) + \mathrm{Re}(A_j)\mathrm{Im}(a) + \mathrm{Re}(B_j)\mathrm{Re}(b) + \mathrm{Re}(B_j)\mathrm{Im}(b) + \mathrm{Im}(\Xi_j), \end{aligned} \qquad (9)$$

where Re(..) and Im(..) are real and imaginary components of a complex value;. $j = 1, N$.

But components y_j and z_j are not independent. In such a case the solution of (9) is determined by the weighted ML procedure [6]:

$$\hat{\boldsymbol{T}} = (\boldsymbol{H}^T \cdot \boldsymbol{W} \cdot \boldsymbol{H})^{-1}(\boldsymbol{H}^T \cdot \boldsymbol{W} \cdot \boldsymbol{Z})$$

where $\mathbf{T}$=[Re(a), Im(a), Re(b), Im(b)]^T is the vector of unknown parameters to be estimated; $\mathbf{Z} = (\hat{y}_1, \hat{z}_1, ..., \hat{y}_N, \hat{z}_N)^T$ are corresponding estimates; H ($2N$×2) and W ($2N$×$2N$) are the matrix of design of experiment and covariance matrix of vector Z respectively

$$\mathbf{H} = \begin{bmatrix} \mathrm{Re}(A_1) & -\mathrm{Im}(A_1) & \mathrm{Re}(B_1) & -\mathrm{Im}(B_1) \\ \mathrm{Im}(A_1) & \mathrm{Re}(A_1) & \mathrm{Im}(B_1) & \mathrm{Re}(B_1) \\ \vdots & \vdots & \vdots & \vdots \\ \mathrm{Re}(A_N) & -\mathrm{Im}(A_N) & \mathrm{Re}(B_N) & -\mathrm{Im}(B_N) \\ \mathrm{Im}(A_N) & \mathrm{Re}(A_N) & \mathrm{Im}(B_N) & \mathrm{Re}(B_N) \end{bmatrix}; \quad \mathbf{W} = \mathrm{diag}(\mathbf{C}, \mathbf{C}, ..., \mathbf{C}); \quad C = (X^T X)^{-1}.$$

and C is the covariance matrix (2×2) of vectors $\dot{u}_j$ defined from (8).

After the estimates of a, b have been found from (9), (10) one can easily calculate the CRC of the load under test $\rho = a/b$. It is important to notice that all the operations performed with primary data (voltages v_{jk}) are linear up to the very last step of calculation of CRC. Hence there were no losses in measuring accuracy caused by "non-optimal" processing of the data obtained. The advantages of the proposed ANA are considered in [7].

The authors carried out the computer simulation of the measurement procedure using the proposed new ANA with four measuring ports. For this purpose the special software package for optimal estimation of load under test parameters using the processing of sampled output voltages of the measuring channels by the ML method was developed.

The dependence of measurement accuracy characterising by the mean squared error upon the signal-to-noise ratio (SNR) at the outputs of ANA's measuring ports is searched. The MR gains A_j, B_j are implied to be exactly known. SNR is equal to b^2/σ^2, where σ^2 is error ξ_{jk} variance in model (6). The measuring process was simulated for the various values of voltage standing wave ration (VSWR) of the load under test. For every specified value of SNR as well as VSWR 10000 simulation measurement experiments of magnitude and phase estimation of the device under test were carried out. After that the mean square error is calculated Figure 3 shows some typical results of simulation presenting the dependences of mean square estimation error of CRC magnitude and phase on SNR at ports outputs for the different values of VSWR of the tested loads. As it can be seen from the curves, the systematic errors of estimation are absent confirming the theoretical conclusion of the precise potential accuracy of the suggested ANA.

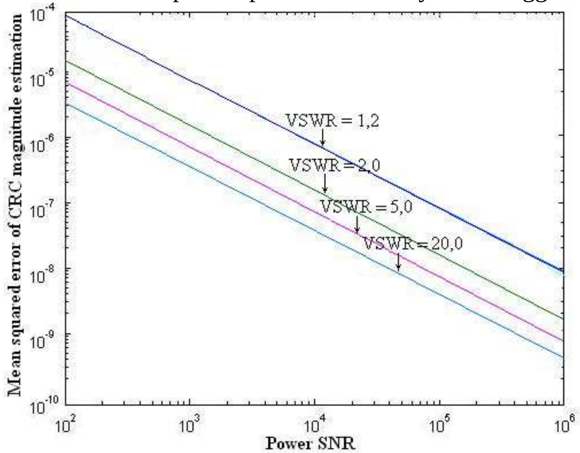

Figure 3. The dependencies of mean squared errors of magnitude estimators on the power SNR for different VSWR

The suggested approach permits the authors to obtain the potentially achievable accuracy of measurements. In addition the necessary amount of precise equipment is drastically dwindled in comparison with conventional ANA based on vector voltme-

ter. Actually, all directional couplers and precision analogue processing circuits are eliminated. Finally, there is no need in large set of precision calibration standards because the proposed ANA can be calibrated using only the set of loads with unknown reflection parameters.

ANA based RFID reader does not suppress spurious signals reflected from the antenna and penetrating the blocking signal from transmitter, and performs conversion of direct additive unlike existing solutions. After that measure amplitude and phase information signal. Therefore, the requirements for methods of modulating the reflected signal tag can be reduced. This leads to a reduction in the cost of tag.

Basing on the above the authors put forward the idea that the proposed ANA based reader of the RFID system may be very promising for use in sorting and stacking production lines.

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